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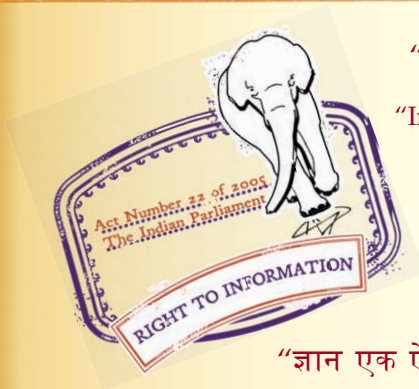
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IS 6134-5 (1980): Methods of measurement of electrical characteristics of microwave tubes, Part 5: Parasitic noise [LITD 4: Electron Tubes and Display Devices]



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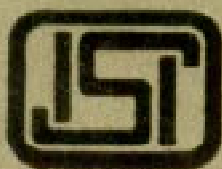
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(Reaffirmed 2003)

Indian Standard

METHODS OF MEASUREMENT ON MICROWAVE TUBES PART V PARASITIC NOISE

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Indian Standard

METHODS OF MEASUREMENT ON MICROWAVE TUBES PART V PARASITIC NOISE

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Indian Standard

METHODS OF MEASUREMENT ON MICROWAVE TUBES

PART V PARASITIC NOISE

0. FOREWORD

0.1 This Indian Standard (Part V) was adopted by the Indian Standards Institution on 11 March 1980, after the draft finalized by the Electron Tubes Sectional Committee had been approved by the Electronics and Telecommunication Division Council.

0.2 This standard (Part V) deals with representative methods for measurement of noise developed in microwave tubes. These methods are summarized and compared in Appendix A to facilitate selection of the best method for use in a particular situation.

0.3 The methods of measurement given in this standard are **sufficiently** accurate to provide a means of assessing noise contributed by the microwave tube within acceptable limits.

0.4 While preparing this standard assistance has been derived from the following publications issued by the International Electrotechnical Commission :

IEC Publication 235-2A (1974)—~~First~~ Supplement to Publication 235-Z (1972) Measurement of the electrical properties of microwave tubes: Part 2 General measurements.

IEC Publication 235-2D (1976)—~~Fourth~~ Supplement to Publication 235-2 (1972) Measurement of the electrical properties of microwave tubes: Part 2 General measurements.

0.5 In reporting the result of a test made in accordance with this standard, if the final value, observed or calculated, is to be rounded off, it shall be done in accordance with IS :2-1960*.

1. SCOPE

1.1 This standard (Part V) deals with representative methods for measurement of noise developed in microwave tubes.

***Rules for rounding off numerical values (*revised*).**

2. TERMINOLOGY

2.1 For the purpose of this standard, the following terms and definitions in addition to those given in IS : 1885 (Part IV/Sec3)-1970* and IS : 1885 (Part IV/Sec6)-1972† shall apply.

2.1.1 Parasitic Noise (in an Electronic Tube) — Energy generated within an electronic tube and detected as undesired sound, visual disturbance or unwanted electrical signal.

NOTE — When the terms 'noise' is used in the following definitions in this document, it is intended to refer to parasitic noise as defined in this clause.

2.1.2 Parasitic Radio-Frequency Noise -- Noise within the radio-frequency band.

3. GENERAL THEORY

3.1 Introduction — Noise measurements are made on microwave tubes to assess the contribution of the tube to system performance degradation arising from noise. Consequently, the type of noise measurement that may be desirable is not solely dependent on the type of tube (that is, oscillator or amplifier) but also on the system application for which it is to be used.

3.1.1 In noise measurements, frequency modulation and phase modulation components are indistinguishable. The result of such a measurement may thus be expressed in either way. Measurements of noise can be made with accuracy only if the system is carefully matched throughout.

3.2 Classification of Noise — In general terms, noise effects may be described as background noise or modulation noise. The difference may be described as follows.

3.2.1 Background Noise — Noise that is always present within a microwave tube. It is very broad-band, having a frequency distribution as shown in Fig. 1. It is usually random noise made up of thermal noise, flicker noise, shot noise, ionic noise and noise arising from any other random process. Background noise in an amplifier is expressed in terms of noise figure.

3.2.2 Modulation Noise — Modulation noise (see also Appendix B) arises from interaction of a carrier with modulating signals (see 3.2.2.1 and 3.2.2.2). This generally results in simultaneous amplitude and frequency modulation sidebands. Modulation noise is measured and expressed in terms of carrier-to-noise ratio. In most practical cases, the measurement may be expressed as the ratio of the carrier power plus overall noise power to the noise power within a stated bandwidth measured at a stated frequency distance from the carrier frequency.

3.2.2.1 Modulation by background noise -When modulation noise is a result of modulation by background noise, the sidebands are correlated with each other but they are not coherent.

*Electrotechnical vocabulary: Part IV Electron tubes, Section 3 Microwave tubes.

†Electrotechnical vocabulary: Part IV Electron tubes, Section 6 Noise in microwave tubes.

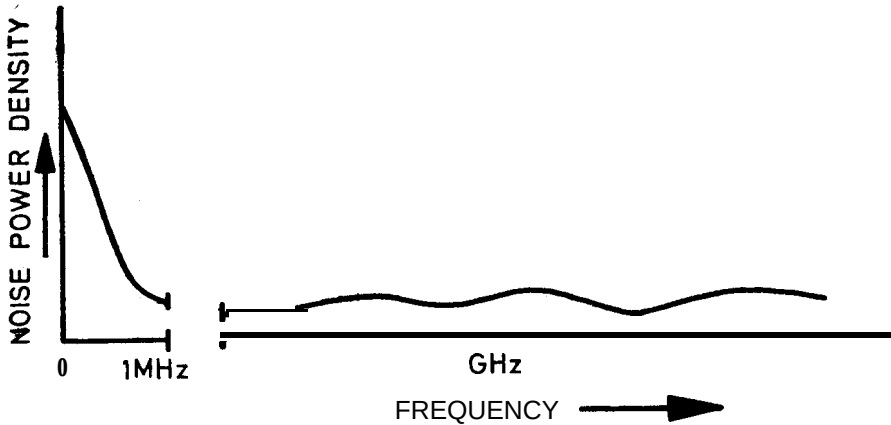


FIG. 1 FREQUENCY DISTRIBUTION OF BACKGROUND NOISE

3.2.2.2 Modulation by other effects— When modulation noise is a result of modulation by a coherent or quasi-coherent signal, such as hum or ionic oscillations, the sidebands are correlated with each other and with the carrier. They are also coherent or quasi-coherent and may give rise to anomalous spectral lines in, for example, doppler radars. The frequency distribution of modulation noise is usually as shown in Fig. 2.

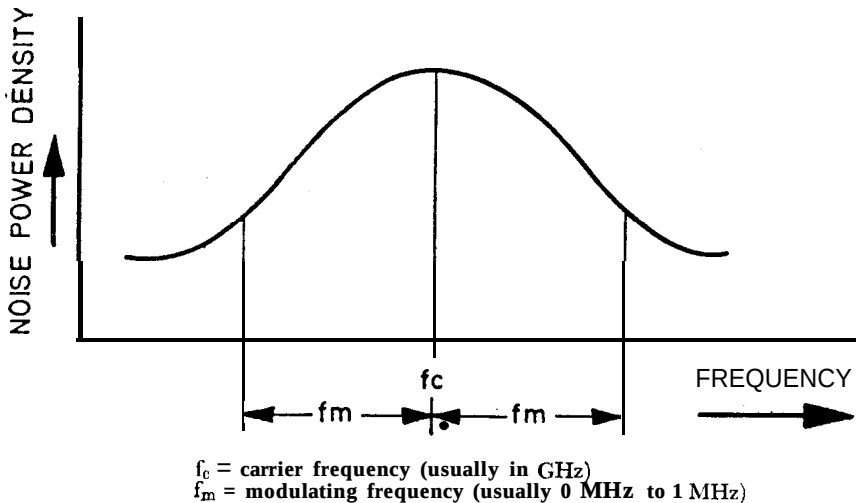


FIG. 2 FREQUENCY DISTRIBUTION OF MODULATION NOISE

4. NOISE REFERENCE AND CALIBRATION SOURCES

4.1 Noise Reference and Calibration Source for Background Noise —

The basic source for noise power in microwave measurements is the calibrated gas discharge tube. The output power from such a tube is very broadband and of the order of 15 dB to 18 dB above thermal noise. The tube is calibrated with a standard for use as a reference for all the usual measurements.

NOTE — In the absence of standard, the manufacturers' test certificate shall be the basis.

4.2 Noise Reference and Calibration Source for Modulation Noise —

In the calibration of the modulation noise measurement system, use should be made of signal generators having internal or external A.M./F.M. modulators of stated low modulation index. In the calibration, the outputs corresponding to several stated values of modulation index are noted and a calibration curve is drawn.

5. MEASUREMENTS

5.0 Introduction

5.0.1 Method of measuring noise are basically similar in that they all involve the measurement of a small amount of noise power in a relatively narrow bandwidth. The availability of various methods of isolation, detection and analysis of this power leads to a wide range of methods for particular applications.

5.02 The methods given in this standard are representative. Other combinations, particularly of calibration and analysis, may give equally useful results for particular applications.

5.0.3 A method commonly used, but not described in detail in this standard, is that of noise factor degradation. This method is more associated with equipment tests than with device tests, but is often used because it represents a real working situation.

5.0.3.1 In this method, the noise factor of a receiver is measured, first with a standard or reference device, and then with a device whose performance is required to be known. The change in noise factor between these two measurements represents the quality of the device being measured, and is expressed as a noise factor degradation. This method is very useful for comparing a series of devices without necessarily giving an absolute value for the device noise factor.

5.0.4 Other effects in the tube being measured, such as unwanted oscillations, instability, drift and jitter, may contribute to the noise power measured, but it is not the purpose of the methods given in this standard to distinguish between the causes of noise, but simply to measure its value.

5.0.4.1 Because of the variety of measurements to be made, it is usually convenient to combine the different methods by use of the same equipment,

which may be connected in such a way that by simple rearrangement several different measurements can be made, such as amplitude modulation noise and frequency modulation noise by changing the phase of a reference carrier.

5.0.5 The methods of measurement described in this standard have been divided into two groups, those related to background noise and those related to modulation noise.

5.0.5.1 Full measurements on a particular type of tube reveal the predominant noise component and subsequently the method of measurement may be restricted to this type of noise.

5.0.5.2 There is, therefore, some overlap between the methods of measurement carried out in a particular sideband frequency range.

5.0.6 The frequency deviation of the signal generator is calibrated by the 'disappearing carrier' method and interpolated to the low modulation indices usually required by reducing the electronic tuning voltage by means of a calibrated variable attenuator.

5.0.7 Precautions are taken to ensure a linear frequency modulation characteristic and to minimize incidental amplitude modulation of the signal generator.

5.0.8 Any effect of unwanted frequency modulation of the calibrating signal generator may be minimized by using a narrow-band-indicator (for example, spectrum analyser) tuned to the frequency of the calibrating modulation.

5.0.9 When a noticeable modulation noise occurs in a region where background noise is present, both methods may be used.

5.0.10 Adequate shielding shall be provided to protect the test equipment against the electromagnetic interference.

5.1 Background Noise Measurements — Background noise in the absence of a carrier signal is generally measured using a heterodyne receiver because direct detection would have poor sensitivity at low power levels and suffer from the very considerable noise generated by a semiconductor detector at very low frequencies. The heterodyne local oscillator provides the correct carrier level for minimum noise factor of the detector. Suitable amplification at intermediate frequency then permits an accurate measurement to be made. When a carrier is present, it is used as a local oscillator signal to transform a portion of the background noise to an intermediate frequency. The bandwidth and centre frequency of the intermediate frequency amplifier determines, which portion of the background noise spectrum in relation to the signal carrier frequency is being measured. This is illustrated in Fig. 3. The level of this portion is measured on an intermediate frequency receiver. The sidebands are then suppressed by appropriate means and a known amount of noise from a noise source is substituted to give the same receiver indication. Thus the intermediate

frequency receiver is used to compare the calibrated noise power with that from the tube being measured.

The advantage of this method is its simplicity and the fact that the determination of the receiver noise factor and calibration of the output meter are necessary.

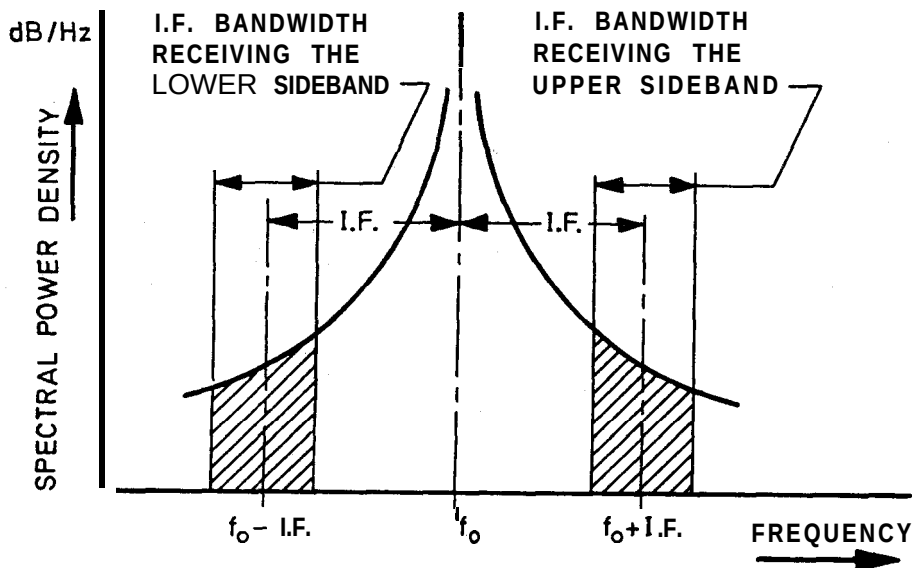


FIG. 3 NOISE SIDEBANDS ASSOCIATED WITH A LOCAL OSCILLATOR IN A RECEIVER

5.1.1 Noise Factor Measurement (Amplifier Tubes) — A block diagram of the circuit for this method is shown in Fig. 4.

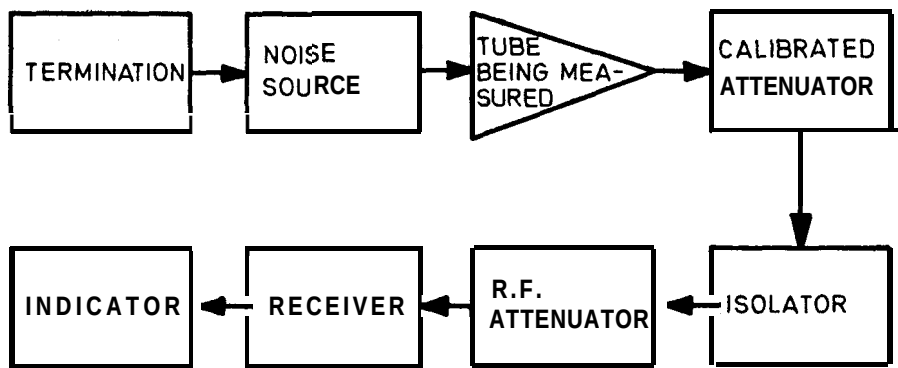


FIG. 4 CIRCUIT FOR THE MEASUREMENT OF AMPLIFIER TUBE NOISE FACTOR

5.1.1.1 Theory — This method is used for measurements in the absence of a carrier for noise factors in the range approximately 3 dB to 30 dB and particularly for travelling-wave tubes. The accuracy can be within 0.1 dB.

The noise output from the amplifier tube is measured with the tube input first connected to a matched termination at the reference temperature and then to a calibrated noise source.

The noise factor is evaluated by comparing the known amount of noise power added from the noise source with the indicated power generated by the amplifier.

The result of the measurement is the measured value of the noise factor averaged over the receiver bandwidth, which usually comprises both upper and lower sidebands equally spaced from the reference frequency by the intermediate frequency of the receiver.

5.1.1.2 Measurement — The tube is operated, under stated conditions, in the circuit shown in Fig. 4. With the noise source switched off, the receiver is tuned to the reference frequency and is adjusted to give a suitable output, which is recorded. With the noise source switched on, the additional attenuation A , required to obtain the same receiver output, is recorded.

When the gain of the tube being measured is sufficiently high and any noise contribution from the attenuator can be neglected, the noise factor F is given by:

$$F = \frac{ENR}{A-1}$$

where

ENR = excess noise ratio of the noise source, and

A = additional attenuation expressed as a power ratio.

The noise factor F may also be expressed in decibels as:

$$F = 10 \log_{10} \frac{ENR}{A-1}$$

NOTE 1 — A heterodyne receiver is usually used. If the receiver is operated over a linear range of intermediate frequency power conversion, an intermediate frequency attenuator may be used in place of a radio-frequency attenuator. The tube noise factor should then be calculated from the formula:

$$F = \frac{F_r - ENR}{A-1} \frac{1}{g}$$

where

F_r = receiver noise factor, and

g = gain of tube being measured (expressed as a power ratio).

NOTE 2 — For a heterodyne receiver, the intermediate frequency and bandwidth should be chosen to represent a situation in which the tube being measured is expected to be wed. For intermediate frequencies greater than 10 MHz, the modulation noise is usually not significant.

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The indicated noise power is the average over the single or double sideband frequency range(s) according to the type of receiver used.

The local oscillator of a double-sideband receiver is tuned to the reference frequency so that the noise power is averaged over two bands equally spaced from the reference frequency by the intermediate frequency. The noise factor over a narrow bandwidth, which will approach a true spot noise factor, can be measured by using a narrow-band receiver of a spectrum analyser.

NOTE 3 — For tubes with a high noise factor, the noise source output may be increased by suitable **amplification**, for example, a travelling-wave tube between the noise source and the tube being measured.

The new value of ENR is then:

$$ENR' = g'(ENR + F') - 1$$

where

g' = noise amplifier gain expressed as a power ratio, and

F' = noise factor of the noise amplifier.

NOTE 4 — The ratio of the noise output power from the tube with the noise source switched on, to the power with the noise source switched off, is known as the Y-factor.

This ratio is given by A as long as the noise contribution from the attenuator and the receiver can be neglected.

NOTE 5 — A continuous reading or automatic noise factor measurement may be made by using a switched noise source and a synchronous detector in the receiver.

The Y-factor is indicated on a meter calibrated directly in noise factor. A suitable switching frequency, for example 1 kHz, permits swept frequency measurements to be displayed on an oscilloscope or recorder.

5.1.1.3 Precautions — The following precautions shall be taken during measurement :

- a) The initial value of attenuation should be as small as possible.
- b) The noise source shall be correctly terminated. Where the termination is not part of the noise tube mount, the manufacturer's requirements shall be complied with.
- c) An isolator may be necessary to prevent interaction with the receiver input circuit.
- d) In double-sideband measurements, it may be necessary to confirm that the noise figure of the tube does not vary substantially over the frequency range represented by the upper and lower sidebands, so that the result obtained may be considered as representing the value at the reference frequency.
- e) If a rejection filter is used to provide single-sideband reception, an isolator should be used to prevent reflection back to the tube being measured.

5.1.2 Sideband Noise Measurement Using a Tunable Filter

5.1.2.1 Theory — A block diagram of the circuit for this method is shown in Fig. 5. The noise sidebands to be measured are associated with a carrier of sufficient power to operate the detector as a mixer with optimum noise factor and conversion loss. The carrier is maintained at this level while the noise in the upper and lower sidebands, corresponding to the

intermediate frequency separation from the carrier and intermediate frequency bandwidth (Fig. 3), is compared with the reference noise power. The comparison is made by suppressing the tube noise sidebands by means of a filter and replacing them by power from the calibrated noise source. Because the intermediate frequency output indication is the average over upper and lower sidebands, noise peaks or symmetry are not indicated.

The main purpose of the measurement is to determine the noise that will be contributed to receiver noise by a local oscillator and hence the intermediate frequency will usually be chosen to lie above any modulation sidebands, but the indication will include amplitude modulation noise plus background noise.

The sensitivity and resolution in low-level measurements depend on the sensitivity of the measuring equipment used.

To attain levels approaching -174 dBm/Hz, it is necessary to use a correlation-level meter or synchronous detector.

5.1.2.2 Measurement — The tube to be measured is connected in the circuit shown in Fig. 5. In this circuit, the tunable filter is used to transmit the reference frequency with or without its associated sidebands. The directional coupler is usually a 3 dB hybrid. With the cavity initially detuned, to transmit the reference frequency only, the variable short circuit is adjusted for minimum transmission, and the cavity is tuned for maximum transmission through the filter.

The tube to be measured is adjusted to operate at the reference frequency under optimum-power conditions. The tunable filter is then adjusted to provide transmission of the reference frequency and its noise sidebands. The power into the detector is adjusted to a suitable level, as indicated by the detector rectified current, and the gain of the amplifier is adjusted to provide a convenient output indication.

The noise source is switched on and the tunable filter is adjusted to provide transmission of the reference frequency only. The attenuator in the noise source arm is then adjusted to return the output indication to the same value as that obtained with the transmission of the reference frequency and the noise sidebands.

It is essential throughout these measurements that the power into the detector be maintained at a fixed level. For that part of the measurement in which the noise source output is not required, sufficient attenuation is inserted to remove effectively the noise source power.

The required noise power P_n per unit bandwidth at the detector input is expressed in dBm/Hz by the formula:

$$P_n = K T_o + ENR - A_{ns} - L_d + S$$

where

$$K T_o = -174 \text{ dBm/Hz};$$

$$ENR = \text{excess noise ratio of noise source, in dB}$$

$$= 10 \log_{10} \frac{T_1 - 290}{290}$$

T_1 = noise temperature of the noise source ;

A_{as} = attenuation of the calibrated variable attenuator, in dB;

L_d = attenuation caused by the coupler and isolator 2, in dB; and

$$S = 10 \log_{10} \frac{S_f}{S_f - 1}$$

where S_f is the sideband suppression factor of the filter expressed as a power ratio greater than unity.

The value of S_f can be determined by a separate measurement of the transmission loss of the filter at the upper and lower sideband frequencies used in the measurement. The loss is measured for the two filter settings used in the measurement. The ratio of these values gives the suppression factor. If the factor varies over the sideband frequency range corresponding to the bandwidth of the intermediate frequency amplifier, an average value shall be taken, and if double-sideband detection is used, transmitted power in the two sidebands for the same incident power shall be added before the ratio is taken.

S_f can also be determined during the measurement by injecting signal at a sideband frequency into isolator 1, *via a* directional coupler. The ratio of the powers required to produce the same additional effect in the receiver in each of the two parts of the measurement gives a value of the suppression factor. It shall be averaged, as for the separate determination.

To determine the carrier-to-noise ratio, the carrier-frequency power is measured at, or referred to, the detector input.

In the preceding expressions, it is assumed that all the circuit elements are in thermal equilibrium at 290 K and that their noise contribution is negligible. In the above method, the noise power of the tube is measurable, provided the value of noise at the detector input does not exceed that of the noise source, which is usually 15 dB or 18 dB above the thermal noise.

If the noise power from the tube is higher than 15 dB to 18 dB, an attenuator may be inserted into the amplifier circuit, for example, between the detector and the receiver amplifier. In this case, arrangements are made so that the attenuator may be switched off during the part of the measurement involving the use of noise source, and it should be ascertained that this does not significantly affect the noise performance of the receiver.

Alternatively, the power from the noise source may be amplified to a level comparable to the noise power being measured. This necessitates measurement of the gain and noise factor of the amplifier used, the amplifier excess noise ratio being expressed as in 5.1.1.2 (Note 3).

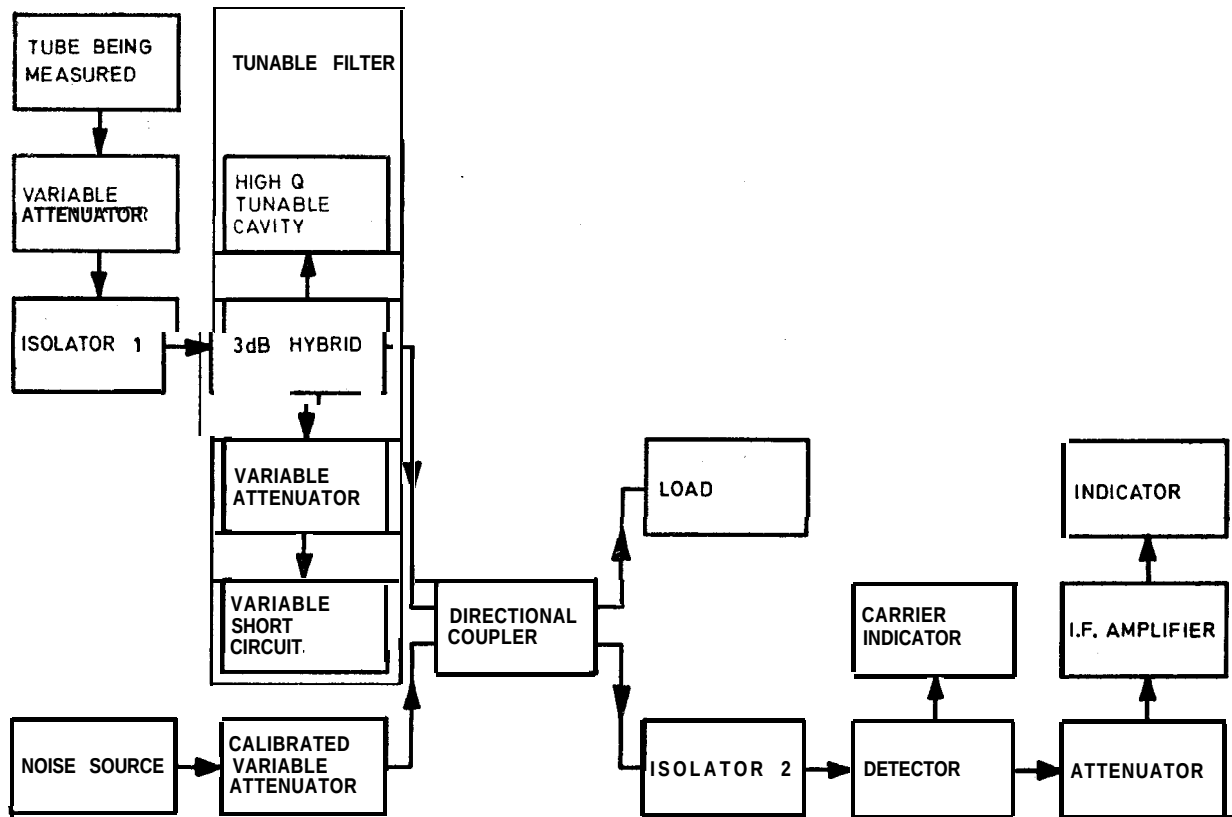


FIG. 5 Circuit Using a Tunable Filter for the Measurement of Sideband Noise

5.1.3 Sideband Noise Measurement (Amplitude Modulation and Frequency Modulation Noises and Their Correlation) Using Two Tunable Rejection Filters

5.1.3.1 Theory—A block diagram of the circuit is shown in Fig. 6. It has two identical microwave rejection filters tuned simultaneously or in turn to the frequencies of the upper and/or lower sidebands measured. The noise in the upper and/or lower sidebands corresponding to the intermediate frequency separation from the carrier and intermediate frequency bandwidth is compared with reference noise power.

The comparison is made by suppressing either or both sideband spectra using rejection filters 1 and 2 and replacing them by power from the calibrated noise source. With the filters detuned, the intermediate frequency output indication is the sum of the powers in the upper and lower sidebands. With both filters detuned simultaneously, the indication is proportional to amplitude modulation noise, that is, in this case the method is capable of indicating the same value as the method according to 5.1.2.

With one rejection filter tuned for suppression of the upper sideband and the other filter detuned, the indication is proportional to the sum of amplitude modulation and frequency modulation noise powers of the lower sideband plus (minus) power dependent on correlation between amplitude modulation and frequency modulation noise. With one filter tuned to the lower sideband the indication is proportional to the sum of amplitude modulation and frequency modulation noise powers of the upper sideband minus (plus) power dependent on the correlation of amplitude modulation and frequency modulation noise.

From the readings obtained with the three modes of filter tuning, one can determine the amplitude modulation noise power, the frequency modulation noise power and their correlation coefficient.

5.1.3.2, Measurement — The tube to be measured is connected in the circuit shown in Fig. 6. In this circuit, rejection filters 1 and 2 are used to transmit the carrier without its associated sidebands or with one (upper or lower) sidebands. To transmit the carrier only, the rejection filters are adjusted for suppression of the upper and lower sidebands. To transmit the carrier with one or both sidebands, the rejection filters are detuned in turn or simultaneously.

The tube to be measured is adjusted to operate at the reference frequency at the maximum power level. The rejection filters are detuned from sideband frequencies. The power into the detector is adjusted to a suitable level and the gain of the amplifier is adjusted to provide a convenient output indication.

The rejection filters are tuned for the suppression of both sidebands, and the noise source is switched on. The calibrated attenuator in the noise source arm is adjusted to return the output indication to the same value as that obtained with the transmission of the carrier and its sidebands. The indication (A_{ns1}) of the calibrated attenuator is then read.

One rejection filter is then detuned and the calibrated attenuator is adjusted to return the indication to the previous value. The indication

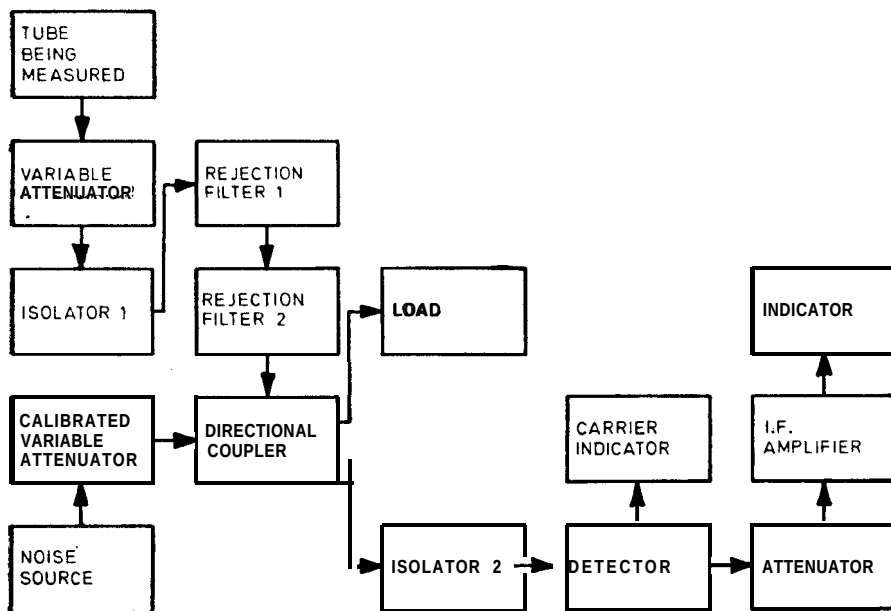


FIG. 6 CIRCUIT USING TWO TUNABLE REJECTION FILTERS FOR THE MEASUREMENT OF SIDEBAND NOISE

(A_{ns2}) of the calibrated attenuator is read. Then the other rejection filter is detuned, and the first one is tuned to the same sideband to which it was tuned previously. The indication (A_{ns3}) of the calibrated attenuator is then read.

NOTE — Amplitude to frequency characteristics of filters used in these measurements are to be identical. Filter bandwidths are chosen so that the suppression of the carrier is negligibly small when the filters are tuned to the required sideband frequencies. When the value of the sideband suppression by the filters within the passband of the intermediate frequency amplifier is greater than 15 dB to 20 dB, the effect of the sideband residual power may be ignored.

From the measurement results, power spectral densities in W/Hz at the detector input are calculated:

- a) The spectral density of A.M. noise power for both sidebands is:

$$P_{A.M.} = \frac{K \cdot T_n}{A_{ns1} \cdot L_d}$$

- b) The spectral density of the upper sideband power is:

$$P_{up} = \frac{K \cdot T_n}{A_{ns2} \cdot L_d}$$

c) The spectral density of the lower sideband power is:

$$P_{\text{low}} = \frac{K \cdot T_n}{A_{\text{ns3}} \cdot L_d}$$

where

K = Boltzmann's constant (1.38×10^{-23} J/K),

T_n = noise temperature of noise source,

L_d = attenuation due to coupler and isolator 2 expressed as a power ratio,

A_{ns1} = attenuation of the calibrated variable attenuator expressed as a power ratio,

A_{ns2} = attenuation of the calibrated variable attenuator expressed as a power ratio, and

A_{ns3} = attenuation of the calibrated variable attenuator expressed as a power ratio.

The spectral density of the frequency modulation noise power, $P_{\text{F.M.}}$, at the detector input (for both sidebands) is calculated from the values obtained according to the formula :

$$P_{\text{F.M.}} = P_{\text{up}} + P_{\text{low}} - P_{\text{A.M.}}$$

and the correlation coefficient C between A.M. and F.M. noise components is defined by the formula:

$$C = \frac{P_{\text{up}} - P_{\text{low}}}{\sqrt{P_{\text{A.M.}}} \cdot \sqrt{P_{\text{F.M.}}}}$$

To determine the carrier-to-noise ratio for A.M. and F.M. components, the carrier-frequency power should be measured at, or referred to, the detector input.

If the noise power from the tube is higher than 15 dB to 18 dB, and attenuator may be inserted into the amplifier circuit, for example, between the detector and the intermediate frequency amplifier. In this case, arrangements are made so that the attenuator may be switched out during the part of the measurement involving the use of the noise source, and it should be ascertained that this does not significantly affect the noise performance of the receiver.

Alternatively, the power from the noise source can be amplified to a level comparable to the noise power being measured. This necessitates measurement of the gain and noise factor of the amplifier used, the amplifier excess noise ratio being expressed as in 5.1.1.2 (Note 3).

NOTE 1 -If background noise is present in the oscillator noise, the method does not allow the separation of background noise from amplitude modulation noise. In this case, the correlation coefficient will not be accurate. However, in any case this method allows the correct measurement of the frequency modulation noise component.

NOTE 2— In this method, the noise power of the tube is measurable, provided the value of noise at the detector input does not exceed that of the noise source, which is usually 15 dB or 18 dB above the thermal noise.

5.1.4 Sideband Noise Measurement Using a Balanced Mixer —

5.1.4.1 Theory — A block diagram of the circuit for this method is shown in Fig. 7. The theory given for 5.1.2 applies, except for the method of suppressing and replacing the tube noise sidebands, and for the case that both detectors of the balanced mixer receive sufficient carrier power.

A balanced mixer is used to suppress noise sidebands and to substitute a known noise power from a calibrated noise source. The principle of this system can be explained as follows :

The balanced mixer comprises a 3 dB hybrid having two radio-frequency input ports (sometimes referred to as signal input and local oscillator input), and a detector in each of the two symmetric output ports. A detected (or intermediate frequency) output is obtained by connecting the two diode detectors of opposite polarity in parallel into a suitable intermediate frequency amplifier. This will be referred to as anti-phase operation.

Used in this way, the balanced mixer provides considerable suppression of unwanted local-oscillator noise sidebands in the intermediate frequency output. These, naturally, have entered the mixer by the same input port as the local-oscillator carrier. The signal, which has entered the mixer by the input port, is unsuppressed in the intermediate frequency output. This is the normal use of a balanced mixer.

If however, the diode detectors used in the mixer are of the **same** polarity, the situation is reversed. Thus, for this mode of operation (referred to as in-phase operation), the signal is suppressed in the intermediate frequency output while the 'local oscillator noise sidebands remain unsuppressed in the intermediate frequency output.

These two modes of operation are used to measure the noise sidebands of an oscillator as shown schematically in Fig. 7. With the tube being measured connected via calibrated level-setting attenuator 1 and isolator to one radio-frequency input port of a balanced mixer, a known noise source is connected via calibrated attenuator 2 to the other radio-frequency input port of the mixer. The intermediate frequency output connections from the diode detectors to the intermediate frequency amplifier are arranged so that the polarity of one of the diodes can be reversed by a switch.

It is therefore possible to switch from in-phase to anti-phase modes of operation of the mixer and compare the tube directly with the known noise source.

In proof of this, let

P_o = excess oscillator noise power per unit bandwidth,

P_n = excess noise power of the noise source and attenuator per unit bandwidth, and

S_f = balanced mixer suppression expressed as a power ratio greater than unity.

Thus in the in-phase mode:

$$\text{I.F. output} = P_o + \frac{P_n}{S_f};$$

In the anti-phase mode:

$$\text{I.F. output} = P_n + \frac{P_o}{S_f}$$

If P_n is adjusted so that the; intermediate frequency output is the same in both modes :

$$P_o + \frac{P_n}{S_f} = P_n + \frac{P_o}{S_f}$$

hence $P_o = P_n$.

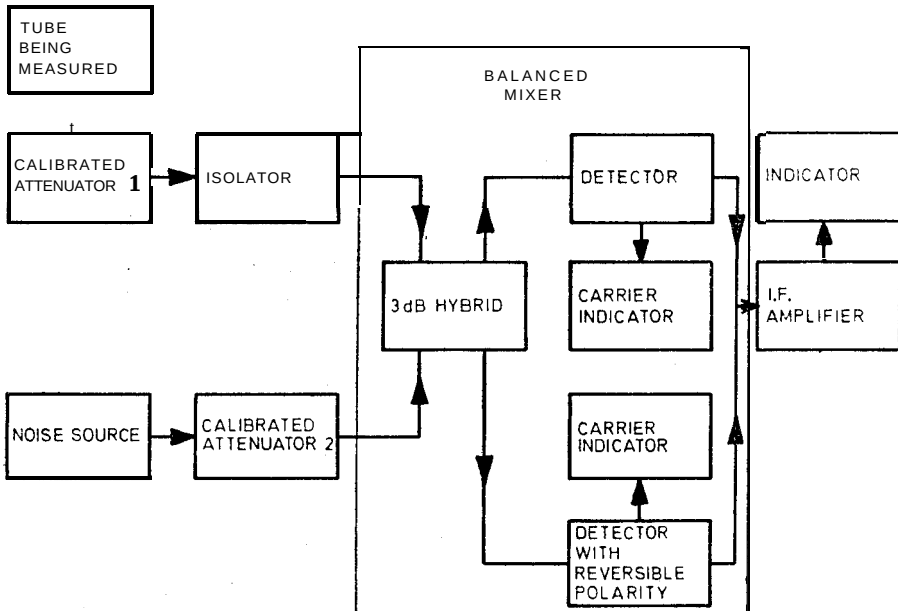


FIG. 7 CIRCUIT USING A BALANCED MIXER FOR THE MEASUREMENT OF SIDEBAND NOISE

5.1.4.2 Measurement — The tube is connected as shown in Fig. 7, and attenuator 1 is adjusted to provide a suitable carrier level at the detectors.

The method of measurement is first to set the intermediate frequency gain so that the level of sideband noise gives a convenient reading on the intermediate frequency output meter when the mixer is switched to the in-phase mode. The mixer is then switched to the anti-phase mode and attenuator 2 is adjusted until the same indication is obtained on the intermediate frequency output meter.

The oscillator sideband excess noise, attenuated by attenuator 1 and the forward loss of the isolator, is then exactly equal to the excess noise provided by the noise attenuated by attenuator 2. The result is expressed in the same units as that of the noise source.

5.1.4.3 Precaution — By techniques of selecting diodes in matched pairs and by applying dc bias to each diode it is possible to obtain power suppression ratios up to 50 dB.

Care shall be taken to arrange the switching of diode intermediate frequency connection so that the intermediate frequency and radio-frequency impedances of the diodes remain unaffected by the switching operation. This is best achieved by maintaining geometrical symmetry in the switching system and arranging identical individual switches for both diodes.

NOTE 1 — It is not necessary to know the amount of suppression provided by the mixer in order to calculate the sideband noise. However, a high degree of suppression provides greater sensitivity of measurement, as may be seen from the proof given above.

NOTE 2 — A further advantage of this method is that optimizing of the suppression ratio does not require auxiliary measurements that could involve dismantling the measurement system. The optimizing can be carried out immediately prior to the measurement at the exact intermediate frequency and radio frequency settings of the actual oscillator under test. All that is required is to introduce maximum attenuation between the noise source and the mixer and maximize the difference between in-phase and anti-phase intermediate frequency output levels by adjustment of dc bias and diode selection.

5.2 Modulation Noise Measurements — In the measurement of amplitude modulation noise, small variations in frequency must not produce variations in the detected signal. This implies a flat frequency response over the measurement band. In the measurement of frequency modulation noise, it is necessary to use a frequency discriminator for detection, and this will result in the detection of both amplitude modulation noise and frequency modulation noise. Means of separating and measuring the two components are described below.

The background noise will inevitably contribute to the measured value but, generally, this is relatively small in amount at frequencies close to the carrier and may be ignored.

At frequencies spaced less than about 1 MHz from the carrier, accurate measurements of amplitude modulation noise are difficult because the video crystal detector contributes low-frequency (flicker) noise of the order of -120 dB/Hz. Account should be taken of the reduction of the conversion

coefficient of the frequency discriminator with reduction of modulating frequency according to the law.

$$\left(\frac{f_{\text{mod}}}{B_{\frac{1}{2}}}\right)^2$$

Where $B_{\frac{1}{2}}$ is half the bandwidth of a discriminator resonator.

The reduction of the conversion coefficient results in an increased importance of detector noises at low frequencies.

NOTE 1 — The use of backward or hot-carrier diodes will reduce the limitation imposed by the flicker noise to a deviation of approximately 10 kHz from the carrier.

NOTE 2 — It should also be noted that the effective bandwidth and incremental sensitivity of the detector and amplifier depend on both the power delivered to the detector and the separate bandwidths of the detector and the amplifier.

The detector is usually operated as a linear detector with a bandwidth much greater than that of the amplifier which follows it, so that the carrier-to-noise ratio is preserved through detection, the direct component of detector output being proportional to the carrier amplitude and the alternating component proportional to the rms amplitude. The carrier-to-noise ratio is given in decibels by the relation:

$$10 \log_{10} \frac{P_c}{P_n} = 20 \log_{10} \frac{V_{dc}}{V_{ac}} \text{ dB, in a bandwidth of } B \text{ Hz}$$

where

P_c = carrier power delivered to the detector (W),

P_n = noise power delivered to the detector (W),

V_{dc} = direct voltage developed across load (V),

V_{ac} = rms noise voltage developed across detector load (V), and

B = amplifier bandwidth (Hz).

The noise voltage may be either mainly at a single frequency or spread over a frequency band. In the latter case, the noise power P_n can be replaced by the noise power density $\frac{P_n}{B}$ and the rms noise voltage by $\frac{V_{ac}}{\sqrt{B}}$, so that the carrier-to-noise density ratio is given in dB/Hz.

Generally, when the ratio is not preserved after detection, it is expressed by the relation:

$$10 \log_{10} \frac{P_c}{P_n} = 20 \log_{10} \frac{V_{dc}}{V_{ac}} R \sqrt{B} \text{ (dB/Hz)}$$

for bandwidth of 1 Hz, where R = Ratio of incremental to direct sensitivity of the detector which depends on detector response to both frequency and power level and on the amplifier bandwidth, which shall be determined by calibration. The noise power may be expressed as a power density.

The technique of carrier cancellation or suppression is useful for measuring high carrier-to-noise ratios when the required sidebands are not too near the carrier and carrier power is well above the minimum required for measurement. Either a narrow stop-band filter or a bridge tuned to the carrier frequency may be used prior to the detector or, particularly in amplifier noise measurements, a sample of the carrier may be reinjected after 'cleaning' if necessary, so as to cancel or reduce the noise carrier and leave the noise sidebands to pass through the detector for measurement. The amount of suppression, in dB, shall be measured and added to the observed ratio, in dB, in order to obtain the carrier-to-noise ratio of the tube being measured.

NOTE 3 -Types of discriminator-Several types of discriminator are available for use in noise measurements. Radio-frequency discriminators are based on either the frequency response of a high-Q cavity or a dispersion characteristic introduced by means of a delay line. The high-Q cavity discriminator is useful for low sideband frequencies in the range 1 kHz (lower with stabilised mean frequency) to 10 kHz and can resolve in a 70 Hz band from 0.01 Hz rms deviation up to 300 Hz. The delay line or two-path discriminator is useful for sideband frequencies in the range 10 kHz to 10 MHz and can resolve in a 3 kHz band from 10 Hz up to 3 MHz rms deviation at a sideband frequency of 10 MHz. Various arrangements to provide carrier suppression or sampling and filtering or reintroduction with phase change can be used, but the basic characteristics of these discriminators are similar near the centre frequency with regard to the form of the curve of output voltage *versus* frequency. Sensitivity to amplitude modulation can be minimized by balancing the carrier component, but as the sideband frequency increases the amount of suppression decreases.

Intermediate frequency discriminators are of either analogue or counter type.

The analogue type is based, as at radio-frequency, on a resonance response, but has the advantages that a limiter can be used to suppress amplitude modulation and that the measurement frequency can easily be changed or swept. A disadvantage is that local-oscillator frequency modulation shall be low compared with that of the tube being measured. It is generally used for wideband frequency modulation measurements for sideband frequencies in the range 300 Hz to 10 MHz and can resolve 5 Hz rms in a 4 kHz band. A counter discriminator is inherently insensitive to amplitude modulation, but is dependent upon local oscillator quality. It is useful for low modulation frequencies particularly in the range where noise and frequency instability are indistinguishable.

5.2.1 Amplitude Modulation Noise Measurement by Direct Detection

5.2.1.1 Theory — A block diagram of the circuit for this measurement is shown in Fig. 8. The power delivered to the detector is adjusted to the value normally required for mixer operation. In this way, the carrier may be considered at the output of a local oscillator, and the noise sidebands as the signal. Provided the combination of detector and first amplifier stage is adequately designed, it is possible to measure within a few decibels of thermal noise level over most of the sideband range from 1 MHz to 100 MHz, to obtain a carrier-to-AM plus background noise

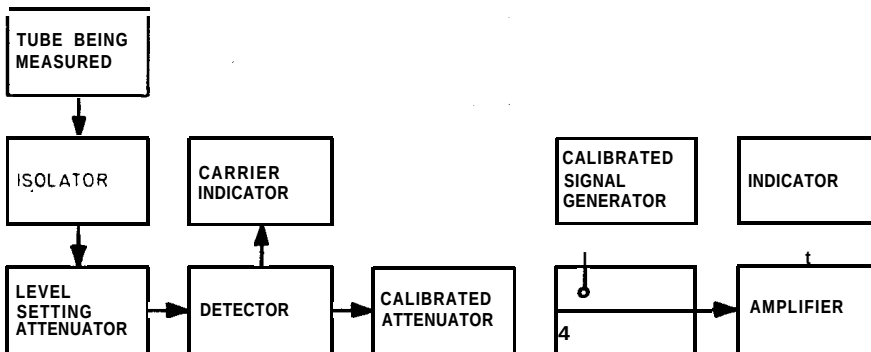


FIG. 8 CIRCUIT FOR THE MEASUREMENT OF A.M. NOISE BY DIRECT DETECTION

ratio of approximately 170 dB/Hz. The method is also useful for noise sideband frequencies down to the carrier where it is used for the measurement of noise caused by power supply ripple and mechanical vibrations, for which a carrier-to-noise ratio of 60 dB to 80 dB may be required.

5.2.1.2 Measurement — The tube is operated under stated conditions. A portion of the output power is applied to the detector, the resulting rectified current being used to indicate carrier level. The carrier is adjusted to a suitable level by means of the level setting attenuator. The alternating components of the detector current caused by amplitude modulation are amplified by the selective amplifier. In order to determine the carrier-to-noise ratio, the output from the amplifier is measured with a rms voltmeter and compared with the power delivered to the detector.

The circuit is calibrated by determining the sensitivity of the amplifier and indicator by means of a calibrating signal generator, and separately determining the power delivered to the detector to provide the reference indication. The calibrating generator may be either a calibrated noise source or a sinusoidal signal generator. In the latter case, it is necessary to plot the amplifier response in order to determine the effective bandwidth. It is also necessary to know the incremental sensitivity of the detector at the sideband frequency so that the changes in power delivered to it can be related to the alternating component measured by means of the amplifier.

The carrier-to-noise ratio can be calculated using the formulae given in Note 2 under 5.2 either for the bandwidth of the selective amplifier or for a bandwidth of 1 Hz.

Calibration may also be carried out by replacing the tube by a signal generator amplitude modulated at known depth and frequency, the modulation frequency being varied to obtain the detector and amplifier bandwidth.

For high power tubes, a portion of the output power is supplied to the detector by means of a suitable directional coupler. The amplifier may be a tunable selective amplifier or a spectrum analyser having a bandwidth that will provide the required resolution of the amplitude modulation sideband components. For the lowest noise level, a synchronous or phase detector should be used as the indicator.

5.2.1.3 Precaution—The video bandwidth of the detector should be sufficient to include all the required noise products.

5.2.2 Frequency Modulation Noise Measurement by Direct Detection (*Simplified Method*) —

5.2.2.1 Theory—A block diagram of the circuit for this method is shown in Fig. 9. A sample of the signal is detected by a discriminator giving an output proportional to both amplitude and frequency. The detected signal caused by amplitude modulation noise is in many cases negligible in comparison with the signal caused by frequency modulation noise and therefore this method can be used to measure frequency modulation noise.

The bandwidth of the discriminator providing the output proportional to frequency should be somewhat greater than the band of measurement but, since the sensitivity is inversely proportional to bandwidth, not much greater than will give reasonable linearity of output versus frequency.

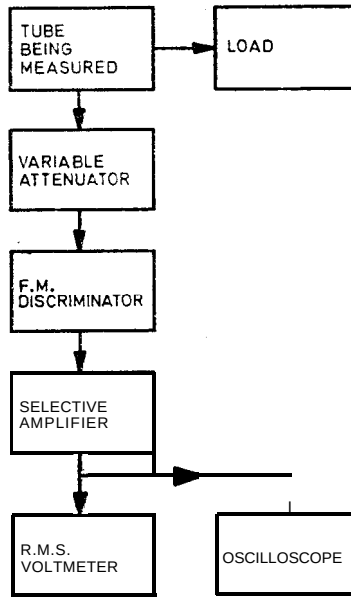


FIG. 9 CIRCUIT FOR THE MEASUREMENT OF F.M. NOISE BY DIRECT DETECTION

The method is also useful for measuring the effects of power supply ripple to within a few hertz from the carrier.

5.2.2.2 Measurement—The tube is operated under stated conditions. One power sample is taken for the measurement circuit which consists of a variable attenuator, a discriminator and a detector. The output voltage of the detector is mainly dc with small alternating components, superimposed proportional to the frequency deviations.

The variable attenuator is adjusted so that the reference dc voltage at the detector is made equal to the optimum value for detector noise factor. The input to the selective amplifier is then proportional to the frequency deviation and to the reference level.

The frequency deviations caused by noise may be measured quantitatively by the calibration of the system using a frequency-modulated signal generator

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in place of the tube being measured. The reading on the voltmeter corresponding to a known amount of frequency deviation provided by the frequency-modulated signal generator is recorded for various deviations and modulation frequencies, and a calibration curve is drawn. The spectrum of the frequency modulation noise can be obtained by tuning the selective amplifier over the required sideband range or displayed by means of a spectrum analyser. A wideband amplifier may be used to obtain the total rms deviation over a stated sideband frequency range.

NOTE—The discriminator used can be a tuned cavity. The tuning frequency and Q are chosen so that the band of measurement is covered by one side of the response curve.

5.2.3 Frequency modulation noise measurement by direct detection using a two-path discriminator.

5.2.3.1 Theory—A block diagram of the circuit for this method is shown in Fig. 10. A two-path discriminator is used because of its good performance over a wide range of sideband frequencies. A balanced detector and differential amplifier provide an output proportional to frequency deviation and relatively insensitive to amplitude modulation. The discriminator bandwidth is chosen to accommodate the required band of measurement on the linear part of the sinusoidal response curve centered on the reference frequency. Excessive bandwidth will reduce the output level due to frequency modulation relative to that due to amplitude modulation. Because of the repetitive nature of the discriminator response at other frequencies, it may be necessary to restrict the bandwidth of the broadband calibrating noise by means of a bandpass filter.

5.2.3.2 Measurement—Prior to the measurement, a signal generator is used to adjust the balanced mixer discriminator together with the differential amplifier, to have the minimum detecting sensitivity for the amplitude-modulation component in order to provide sufficient suppression of that component. The output from the selective amplifier is then assumed to be proportional only to the frequency deviation.

The tube is operated under stated conditions. The power drive of the detectors is adjusted to the value normally required for mixer operation, and is maintained constant throughout the measurement. The carrier power is measured at, or referred to, the input of 3 dB hybrid 3.

The frequency modulation component of noise power is evaluated in a manner similar to that for the measurement of sideband noise, using a tunable filter to transmit the carrier frequency with or without its associated noise sidebands (see 5.1.2 and Fig. 5).

The noise sideband spectrum can be measured by tuning the selective amplifier or by means of a spectrum analyser. For narrow-band measurements, the calibration can be carried out using a signal generator with calibrated frequency modulation.

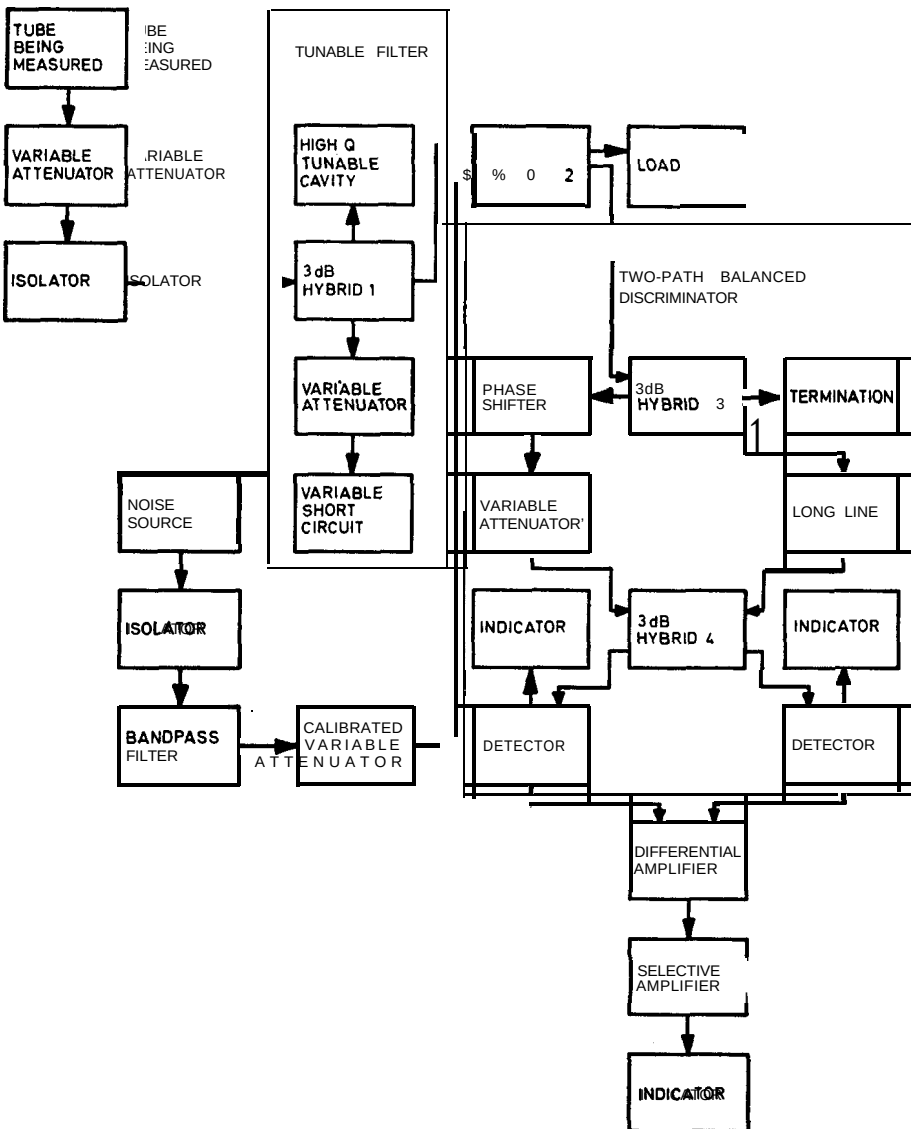


FIG.10 CIRCUIT USING A TWO-PATH DISCRIMINATOR FOR THE MEASUREMENT OF F.M. NOISE BY DIRECT DETECTION

5.2.3.3 Precautions— The following precautions should be taken during measurements.

- a) **3 dB** hybrids 3 and 4 shall be carefully balanced.
- b) The semiconductor diodes used as detectors shall be a matched pair over the dynamic range expected.
- c) The bandwidth of the detectors and the differential amplifier should exceed the frequency range of the noise sideband to be measured.

5.2.4 Amplitude and frequency modulation noise measurement close to the carrier using a high-Q microwave discriminator.

5.2.4.1 Theory—A block diagram of the circuit for this method is shown in Fig. 11. Measurements of frequency modulation noise components at frequencies spaced less than about 1 MHz from the carrier become less accurate when direct detection methods are used (see 5.2.3) and consequently measurements are made in that sideband range by use of a discriminator at intermediate frequencies, often after carrier suppression at the radio frequency.

In the case of measurement at the intermediate frequency, the sensitivity of the discriminator is limited by the overall noise factor of the mixer and intermediate frequency amplifier, together with the limited power of the signal input that can be applied to a microwave mixer. Because of these difficulties, it is usually easier to use a microwave discriminator at the radio frequency.

In operation at the radio frequency, however, the input is similarly limited if the discriminator consists of a detuned resonant cavity. This limitation can be overcome if the microwave discriminator is balanced, that is, if the input carrier power is suppressed before the signal is applied to the mixer crystal, and then later reintroduced at a suitable level and in the correct phase to provide detection of the required modulation noise sideband.

This can be achieved at the intermediate frequency where the reference carrier can be used for the switching signal in a phase detector. An advantage of this method is that the requirement for a very stable local oscillator is removed. By the use of an additional phase detector switched in quadrature with the first, a simultaneous measurement of amplitude modulation and frequency modulation can be made so that their correlation can also be measured. The method described here uses only one phase detector.

The sensitivity of this method increases with the modulation frequency, and is considerably higher than that obtained by use of direct detection methods at the higher modulation frequencies.

In this method, a sinusoidal variation of signal frequency is first changed

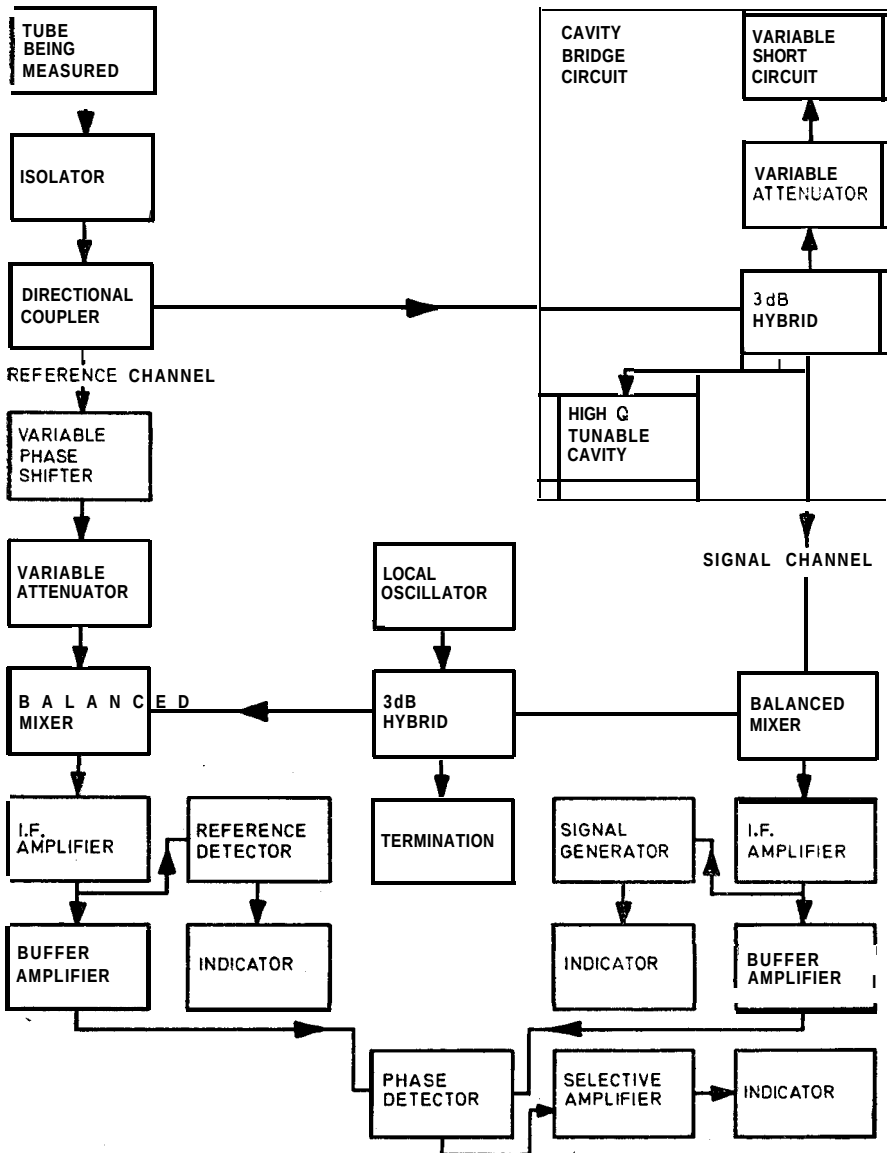


FIG. 11 CIRCUIT FOR THE MEASUREMENT OF A.M. AND F.M. NOISE CLOSE TO THE CARRIER FREQUENCY BY A HIGH-Q MICROWAVE DISCRIMINATOR

in phase in proportion to the deviation, then transposed to intermediate frequency with the phase change preserved and finally applied to a phase detector to yield an output proportional to frequency.

The following property of a resonant cavity is used to obtain the change in phase and at the same time to suppress the carrier.

When a modulated signal is applied to a resonant cavity, tuned to resonance at the carrier frequency, the signal reflected from the cavity depends upon a known relationship. The in-phase and quadrature components of the voltage reflection factor near resonance are shown in Fig. 12.

The reference plane is that at which the cavity presents a short circuit at frequencies well off resonance, at which the cavity is matched, and only the quadrature component is significant, having a reflection factor n , given by:

$$n = \frac{(f - f_0)}{f_0} 2Q_L$$

where

f_0 = resonance frequency (a carrier frequency in the measurement),

f = a sideband frequency near the resonance frequency, and

Q_L = loaded Q .

The effect on a carrier modulated at single low frequency, for example, is to suppress the carrier, change the phase of one of the sidebands, and change the phase of the other by the same amount plus 180°. Because of the high degree of suppression, the power delivered to the measuring circuit may be very much higher than with direct detection, thus considerably increasing the sensitivity to sideband noise. The sidebands then pass through the signal channel and are detected by the phase detector. Depending on the phase of the reference signal, either amplitude modulation or frequency modulation sidebands will be detected.

The varying output from the phase detector is examined by means of a selective amplifier or spectrum analyser, and the magnitude of the sideband components measured. The measured output from the reference channel represents the carrier amplitude, and hence the ratio between these two quantities will provide the required carrier-to-noise ratio.

The effect of local oscillator noise near the signal frequency is made negligible by the choice of oscillator and the use of a balanced mixer that can provide a 10 dB to 20 dB improvement. The amplitude modulation noise near the local oscillator frequency may be ignored provided that leakage power from the discriminator is small compared with the local oscillator power at the mixer.

Frequency (or phase) modulation of the local oscillator is insignificant if the signal and reference intermediate frequency amplifiers have equal slopes

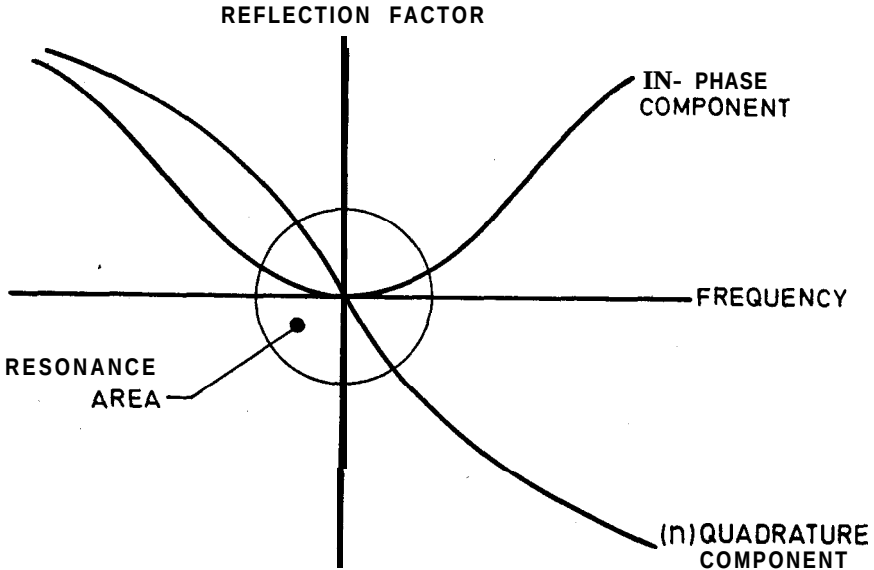


FIG. 12 REFLECTION FACTOR OF CAVITY NEAR RESONANCE

or time delays, since the effect would produce equal phase modulation of reference and leakage carrier. With a discriminator balance of 70 dB, the local oscillator frequency modulation noise may, in terms of frequency deviation, be 10^4 times worse than that of the signal being measured, before causing appreciable error. At frequencies well outside the cavity passband, the discriminator simply becomes a carrier filter, the noise sidebands being reflected with undiminished amplitude.

The circuit basically comprises an input waveguide terminated by a matched resonant cavity. Coupled to the input waveguide are two channels : a signal channel and a reference channel. In the signal channel, the reflected signal from the cavity is coupled by means of a 3 dB hybrid to a mixer followed by an intermediate frequency amplifier, and finally a phase detector. A power sample from the tube being measured is coupled via the reference channel to the phase detector.

In this circuit, the resonant cavity forms part of a bridge circuit incorporating a variable attenuator and an adjustable short circuit, which provide a variable impedance on one arm of the hybrid so that the cavity impedance at resonance may be balanced and the minimum carrier power transmitted to the signal channel. The amount of attenuation in the balancing arm for the optimum balance condition also provides a useful indication of the match presented by the cavity, and this fact may be used to ensure that the

cavity has not been tuned in error to any spurious modes that may exist within the cavity tuning range.

The carrier power is normally dissipated in the cavity, which may require cooling for high power. Since the sensitivity increases with carrier power, the highest level consistent with achievable balance should be used. A coupler or an attenuator may be used to reduce the power.

Amplitude modulation noise occurring similarly in the two channels is suppressed at the output by the balanced action of the phase detector provided the sidebands do not extend appreciably towards the frequencies at which the quadrature components maximize. A rejection value of more than 40 dB is not difficult to achieve.

NOTE 1—To measure sidebands up to 120 kHz requires a cavity 3 dB bandwidth of about 600 kHz.

NOTE 2—For high input powers applied to the discriminator and for high degree of carrier frequency suppression the measurements become complicated if the tube being measured has poor long-term stability. For example, in operation at a carrier frequency of 10 GHz with a discriminator bandwidth of 600 kHz, a frequency stability of about 10^{-6} is required for the measurement to be carried out conveniently, provided that an input level of the order of 1 W is used.

NOTE 3—If the tube is not sufficiently stable, an automatic frequency control system should be used to lock together the frequencies of the discriminator and the tube being measured.

5.2.4.2 *Measurement (FM)*—The equipment is set up and calibrated prior to the measurement by connecting a signal generator to the measuring circuit input. An unmodulated carrier at the reference frequency is used to adjust the cavity bridge circuit for maximum carrier suppression in the signal channel. The local oscillator frequency is set to centre the intermediate frequency in the intermediate frequency amplifier band. With the correct level in the reference channel, the carrier is amplitude modulated and the reference phase is adjusted for cancellation in the phase detector output. The carrier is then frequency modulated by a known amount and the output indicator calibrated.

The tube is next connected in place of the generator and operated under stated conditions at precisely the reference frequency. The power is adjusted to provide the correct level in the signal channel. As a result of carrier suppression, a power of several hundred milliwatts may be required; lower power simply reduces the measurement sensitivity. The reference channel power is also adjusted to the correct level.

The frequency modulation spectrum components are then measured and referred to the carrier power to provide the carrier-to-noise ratio, or to thermal noise level in effective bandwidth of the selective amplifier. The measured noise includes background noise.

5.2.4.3 *Measurement (A.M.)*—A lower sideband frequencies, the reference channel can be used alone to measure amplitude modulation

plus background noise in a similar manner to direct detection. At higher sideband frequencies, the carrier suppression obtained by using both channels permits a more sensitive measurement.

The reference channel phase shifter is adjusted to a phase that cancels the frequency modulation component, and the amplitude modulation sideband noise is measured as for frequency modulation (see 5.2.4.2).

The sensitivity of measurement increases with sideband frequency at 6 dB per octave and directly with input power. Therefore, with a limiting sensitivity of 110 dB carrier-to-noise ratio at 1 W input 1 kHz from the carrier, the best method may be determined.

5.2.4.4 Precautions — The local-oscillator level-setting attenuator of the terminations in both channels shall present a good match to the waveguide so that interaction between the two channels is reduced to a minimum.

In the measurement of very small values of frequency modulation noise, it is essential that the ripple content of power supplies at the relevant sideband frequencies be very low; for example, a tube with a frequency pushing figure of 2 MHz per ampere, fed from a power supply having a ripple content of $10\mu\text{A}$, would produce 20 Hz deviation.

Care is necessary in regions near the edge of the cavity resonance band because of changing phase and sensitivity. For the highest sensitivity very near the carrier, it may be necessary to maintain cavity tuning by adjustment during the measurement.

5.2.5 *Measurement of Noise in Microwave Amplifiers Using a Microwave Phase Discriminator*

5.2.5.1 *Measurement of c. w. amplifiers*

Theory — A block diagram of the circuit arrangement used for this method is shown in Fig. 13.

A sample of the amplified signal is compared with a sample of the signal driving the amplifier. By adjustment of the attenuation and phase in the reference channel, the two carriers compensate within the 3 dB hybrid. The signal then appearing at port (3) consists only of the noise sidebands introduced by the amplifier and its power supply.

The noise sidebands are demodulated in a low-noise mixer, such as a Schottky diode, by reinjection of the carrier at low level, typically of the order of 1 mW.

Cancellation of the carriers in the 3 dB hybrid can be of the order of 40 dB. It is then possible to operate the phase discriminator at a relatively high level, typically 100 mW to 1 000 mW. Consequently, the sidebands appearing at port (3) of the hybrid will correspond to a 100 mW to 1 000 mW carrier whilst the mixer will receive less than 0.1 mW (40 dB directivity) of this carrier. By varying the phase of the carrier reinjected to the mixer, it is now possible to measure separately the noise components

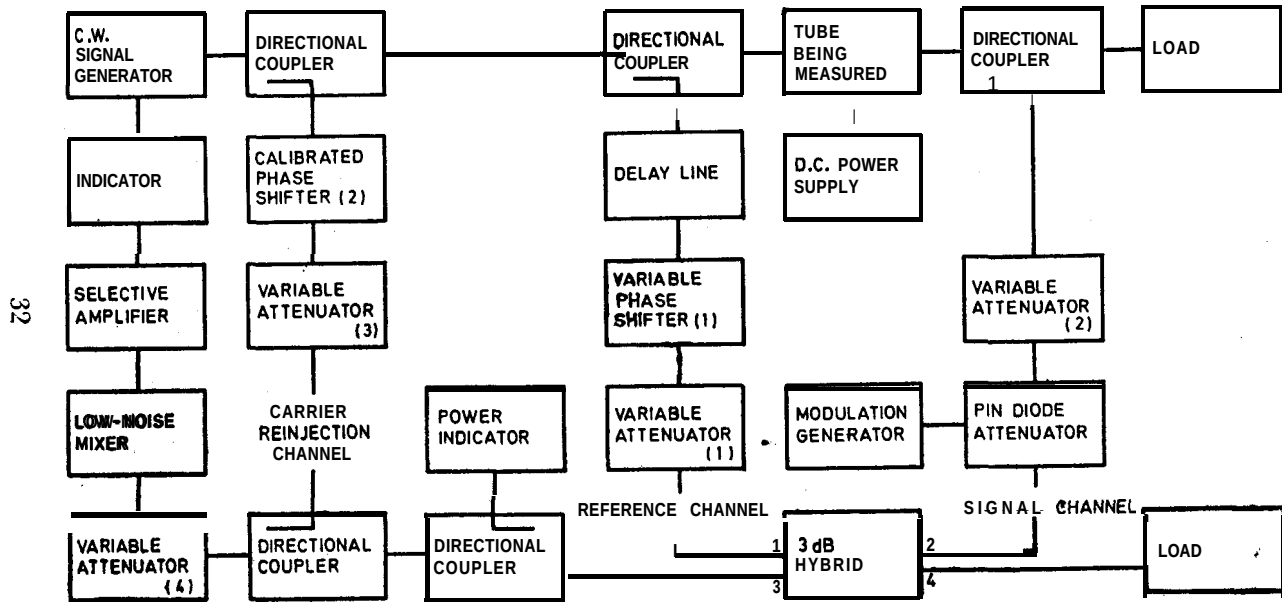


FIG. 13 CIRCUIT FOR MEASUREMENT OF NOISE IN A C.W. AMPLIFIER USING A MICROWAVE PHASE DISCRIMINATOR

in phase or in quadrature with the carrier, corresponding respectively to A.M. and F.M. noise.

Calibration is performed by introducing a known amount of amplitude modulation on one of the channels by the use of a calibrated PIN diode attenuator driven by a modulation generator. The calibrated phase shifter in the carrier **reinjection** channel is adjusted for maximum reading on the indicator. The equipment is now set for A.M. noise measurements. For F.M. noise measurements, it is only necessary to change the phase of the reinjected carrier by 90° , the calibration remaining valid.

A delay line can be used in the reference channel to equalize the electrical lengths of the two channels, including the amplifier, and to minimize the frequency sensitivity of the phase discriminator balance.

The sensitivity of this type of equipment is limited by the **noise** factor of the receiver (for example, low-noise mixer and selective amplifier) and the total rf input to the mixer. If a conventional point contact diode is used, the noise figure of the receiver at X band, which can be of the order of **10 dB** at 10 MHz intermediate frequency, increases linearly to **50 dB** at **1 kHz**. By using a Schottky diode, the noise figure remains constant down to a few kilohertz from the carrier (typically **5 kHz**) and then increases at a rate of **10 dB** per decade.

If a **10 dB** receiver noise factor is achieved with **1 mW** total rf input to the mixer (**0 dBm**), the minimum detectable signal will be **-134 dBm/kHz** (minus **144 dBm/kHz**, thermal noise at ambient temperature, plus **10 dB** noise figure). The phase discriminator being balanced at a power level **20 dB** to **30 dB** higher than the local oscillator level, the theoretical limit of the carrier-to-noise measurement is **154 dB/kHz** to **164 dB/kHz**. In fact, the practical sensitivity is limited by the noise of the signal generator present in the reinjected carrier and the minimum detectable noise contribution of the amplifier is **20 dB** to **30 dB** below the signal generator noise.

NOTE -It is also possible to use a superheterodyne receiver with two balanced mixers to ensure the transfer of the noise to be measured to an intermediate frequency. In this case, a noise figure of the order of **10 dB** can be obtained by using conventional point contact diodes. A block diagram of the receiver is shown in Fig. 14.

Both **A.M.** and **F.M.** noise are determined in terms of the carrier-to-noise ratio. If one wishes to express **F.M.** noise in terms of carrier frequency deviation, the following relation is used:

$$\Delta F_{\text{rms}} = \frac{f}{10^{A/20}} \quad (\text{two sidebands})$$

where ΔF_{rms} is the rms deviation of the carrier frequency at modulating frequency f , corresponding to a carrier-to-noise power ratio equal to A decibels, in the bandwidth of measurement.

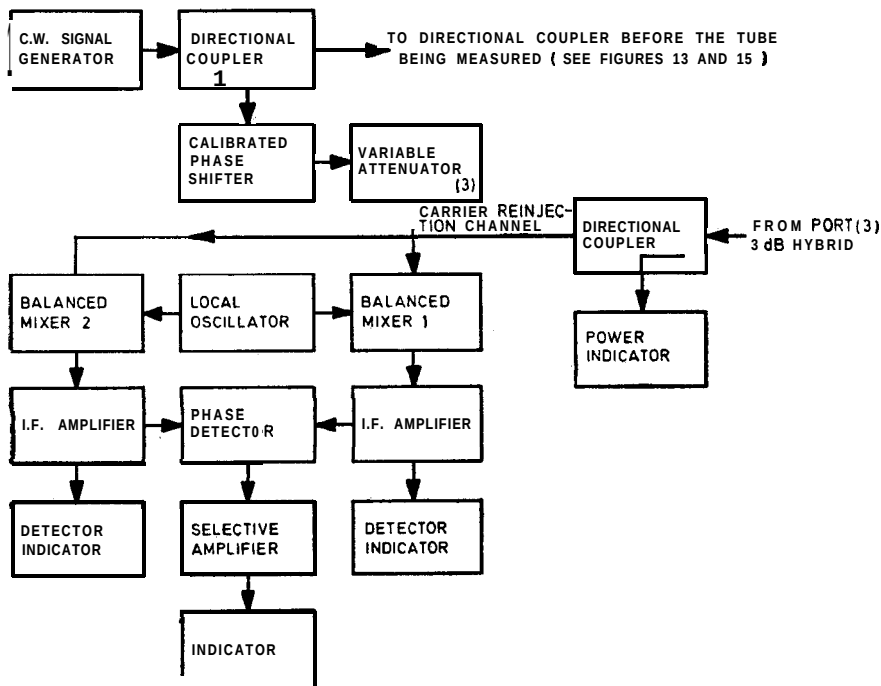


FIG. 14 BLOCK DIAGRAM OF THE SUPERHETERODYNE RECEIVER FOR THE MEASUREMENT OF NOISE IN A MICROWAVE AMPLIFIER

Measurement — The low-noise mixer is protected from overload by setting variable attenuator (4) at maximum attenuation (30 dB minimum). The variable attenuator (1) is set to maximum attenuation (no reference channel power). The amplifier is operated under stated conditions. The signal channel level is set by variable attenuator (2) to an appropriate value, 100 mW to 1 000 mW, read on the power indicator.

The phase discriminator is balanced in successive approximations by actions on variable attenuator (1) and phase shifter (1). Indication of balance is given by near zero reading on the power indicator. When balance is achieved variable attenuator (4) is set to minimum. Carrier reinjection level is set to an appropriate value (1 mW) by variable attenuator (3). A known amplitude modulation of low index is introduced by the modulation generator driving the PIN diode attenuator. The selective amplifier is tuned to the modulating frequency.

The calibrated phase shifter is adjusted for maximum reading on the

indicator. By tuning the selective amplifier, it is now possible to measure the A.M. noise power in the selective amplifier bandwidth as a function of frequency deviation from the carrier.

For F.M. noise measurements, the calibrated phase shifter setting is changed by 90° .

Precautions — The following precautions shall be taken during measurements :

- a) Any appreciable imbalance of the phase discriminator resulting for instance from switching off the dc power supply of the amplifier tube during noise measurements will cause the mixer diode to burn out. During any adjustment of the tube input, such as change of operating voltage or frequency, it is necessary to protect the mixer by setting variable attenuator (4) to maximum attenuation.
- b) Known amplitude modulation used for calibration must be of sufficiently low index to ensure good linearity of the mixer, selective amplifier and indicator from calibration down to noise levels.
- c) The indicator should read the rms value of the amplified noise voltage.
- d) If a spectrum analyzer is used to give a visual indication of noise, it should be noted that this cannot be used for quantitative measurements.
- e) Care shall be taken to ensure that the PIN diode modulator does not introduce unwanted phase modulation and noise.

5.252 Measurement of pulsed amplifiers

Theory -A block diagram of the circuit arrangement used for noise measurements on pulsed amplifiers is shown in Fig. 15.

If a C. W. carrier is 100 percent amplitude modulated by rectangular shaped pulses of duration t at a pulse repetition. Frequency f (pulse repetition period $T = \frac{1}{f}$), the transmitted signal appears in the frequency domain as a spectrum consisting of an infinite number of discrete lines separated by a constant frequency interval equal to f hertz, the amplitude of which is maximum at the carrier frequency and varies as:

$$\frac{\sin \pi \cdot \Delta f \cdot t}{\pi \cdot \Delta f \cdot t}$$

where Δf is the deviation in frequency from the carrier and t is the pulse duration.

The main lobe of the spectrum is then $\frac{2}{t}$ Hz wide and will comprise $\frac{2}{\Delta f}$ spectral lines.

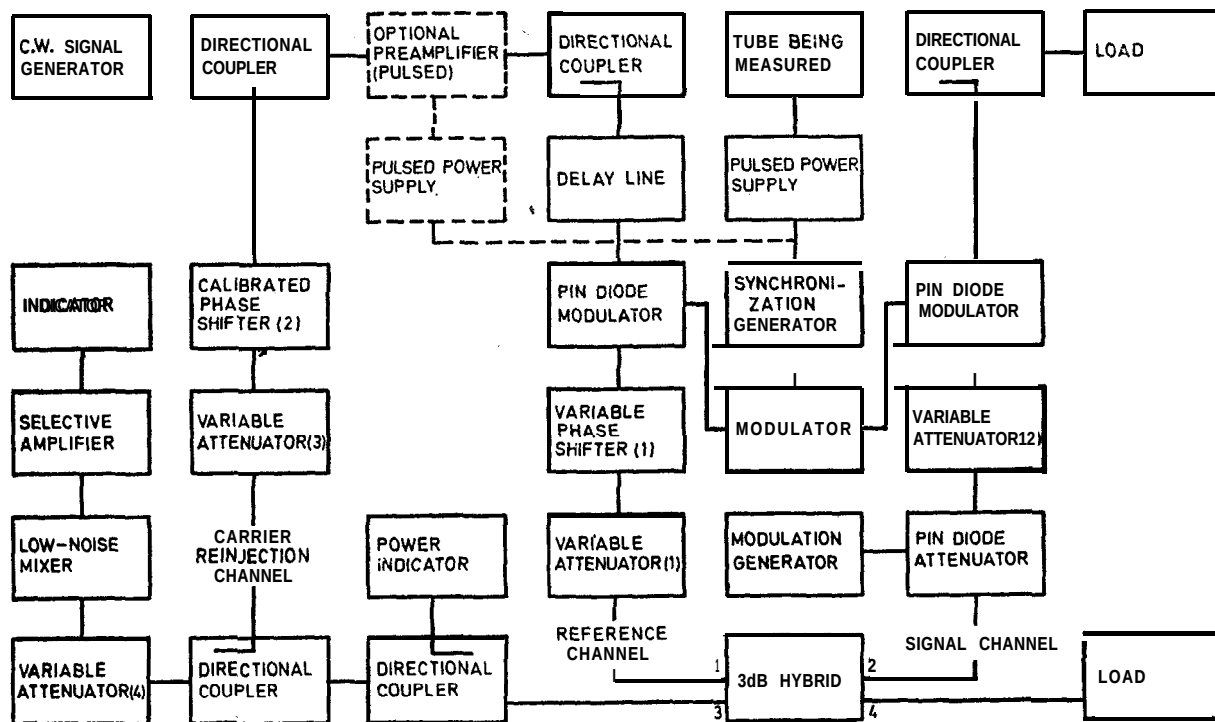


FIG. 15 CIRCUIT FOR MEASUREMENT OF NOISE IN A PULSED AMPLIFIER USING
4 MICROWAVE PHASE DISCRIMINATOR

If P_o is the peak transmitted power of the unmodulated carrier, the mean or average transmitted power is:

$$P_m = P_o \cdot \frac{t}{T} = P_o \cdot t \cdot f$$

and the power contained in each of the main spectral lines (carrier or close to carrier) is:

$$P_e = P_o \cdot \left(\frac{t}{T} \right)^2 = P_o \cdot t^2 \cdot f^2$$

The circuit used is basically the same as that used for C. W. amplifiers except that the balance of the phase discriminator must be achieved under pulsed conditions, which requires that the pulse shapes should be exactly the same in reference and signal channels. This is achieved by the use of two identical PIN diode modulators, one in each of the two channels, driven from the same pulsed voltage generator.

Each PIN diode modulator, normally biased for maximum attenuation, is pulsed to minimum attenuation during a fraction of the total of the flat top portion of the pulse of the amplified signal. The two pulsed signals, being made identical in shape, can be balanced in the phase discriminator.

As pulse amplifiers are generally operated at high peak power output with limited gain, it is sometimes necessary to add a preamplifier between the C. W. signal generator and the tube being measured. This preamplifier can be pulsed.

The carrier reinjection shall be provided by a C. W. operating signal generator.

The power in the reference and signal channels is limited by the peak power handling capability of the PIN diode modulators, usually 1 000 mW.

The selective amplifier must have enough selectivity to allow noise measurements between two spectral lines.

The sensitivity of this type of noise measurements is limited by the noise spectrum of the signal generator because of imbalance in the phase discriminator. As the noise spectral power density in oscillators generally decreases very rapidly when the frequency deviation from the carrier increases, the sensitivity is strongly dependent upon the frequency interval between spectral lines (pulse repetition frequency).

At 1 000 Hz pulse repetition frequency a sensitivity of 70 dB in 1 kHz bandwidth below one of the main spectral lines can be achieved.

NOTE — There is a definite advantage in not performing the measurement around the carrier spectral line but preferably farther from the carrier frequency, as low-frequency noise contributions from the mixer and signal generator are then reduced.

Measurement — The method of measurement shall be the same as given in **5.2.5.1**.

5.2.6 Modulation Noise Measurement Using a Swept-Frequency Receiver or Spectrum Analyser

5.2.6.1 Theory-A block diagram of the circuit for this method is shown in Fig. 16. The method is useful for the continuous display and measurement of modulation sideband power or undesired oscillation such as may occur in backward-wave oscillators. The distribution of power in the frequency range of measurement is analysed by transposing the carrier and its sidebands to a suitable intermediate frequency and examining, by means of an intermediate frequency band-pass amplifier, the power in a part of the intermediate frequency spectrum selected by tuning or sweeping the local oscillator. An intermediate frequency bandwidth from 1 kHz to 2 MHz is commonly used and, with a carrier power of about 0.1 mW at the mixer, sideband power from -20 dBm to -90 dBm may be measured. A sweep range of up to 2 GHz can be achieved. Background noise is not usually significant in this measurement.

The intermediate frequency should be high enough to avoid image channel responses when the required rf range is swept by the local oscillator. Thus a high intermediate frequency permits a greater sweep range and greater panoramic display for measurement.

The video signal represents the average power within the intermediate frequency band which should be narrow enough to resolve any noise peaks that are required to be measured. Modulation frequencies lower than the intermediate frequency band cannot be resolved.

The local oscillator noise may contribute to the indicated noise, but this can be minimized by correct operation of the mixer and choice of a local oscillator that can be phase locked. The sideband power thus displayed may include both amplitude modulation and frequency modulation. In some cases, for example backward-wave oscillators, the amplitude modulation may be small and the sideband power considered entirely due to frequency modulation.

5.2.6.2 Measurement-The tube is operated under stated conditions in the circuit shown in Fig. 16 and a suitable power is delivered to the swept receiver. The levels of sideband peaks are observed by varying the local oscillator frequency at a suitable sweep rate and displaying the responses due to the carrier and sidebands. The receiver swept-frequency range is adjusted to include as many as possible of the sidebands to be measured and, if required, the tube being measured is tuned over a range of frequencies so that variations of sideband power can be observed. The sideband levels are measured relative to the carrier by means of the radio frequency and intermediate frequency variable attenuators or the oscilloscope display, if calibrated. The frequency separation from the carrier, of each peak or

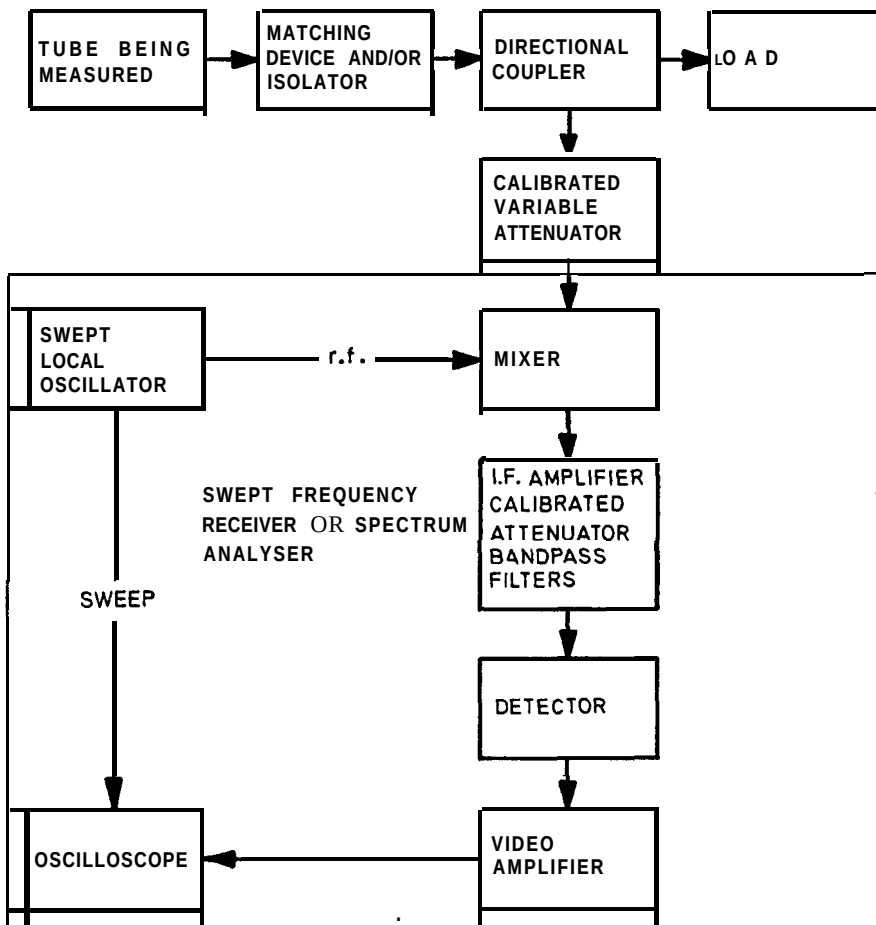


FIG. 16 CIRCUIT FOR THE MEASUREMENT OF MODULATION NOISE USING A SWEEP-FREQUENCY RECEIVER OR SPECTRUM ANALYSER

of the centre frequency of each noise band is noted. The carrier-to-noise ratio is given by the formula:

$$\left[\frac{C}{N} \right] = 10 \log_{10} \frac{P_c}{P_s} \text{ dB, at modulation frequency } f_m$$

$$= (R_c - R_s) \text{ dB}$$

where

P_c = carrier power (W)

P_s = sideband power (W),

R_c = carrier level (dB), and

R_s = sideband level (dB).

The resolution of the measurement is limited by the intermediate frequency response. Therefore, if the bandwidth of the sideband response is approximately the same as that of the carrier, the ratio should be expressed as above; if it is wider, the ratio should be expressed as carrier-to-noise power density, given by

$$(R_c - R_s + 10 \log_e B) \text{ dB/Hz}$$

where B—intermediate frequency bandwidth (Hz).

If the modulation noise is known to be frequency modulation noise, the deviation can be expressed as:

$$\Delta f = 2f_m \sqrt{\frac{P_s}{P_c}} = 2f_m \left[\text{antilog}_{10} \left(\frac{R_c - R_s}{20} \right) \right]$$

where

Δf = peak deviation (Hz), and

f_m = modulation frequency (Hz).

It should be noted that P_s represents the power in one sideband only. For pure frequency modulation of low modulation index, there will be one equal image sideband and either sideband can be taken for measurement of P_s . If both amplitude modulation and frequency modulation are present at the same modulation frequency, the sidebands may have different power levels according to the phase of amplitude modulation relative to frequency modulation. It is not possible to separate the amplitude modulation and frequency modulation components without a separate measurement to determine one or the other.

If the sideband is wide, the frequency modulation noise is expressed as an rms deviation per Hz, thus

$$\Delta f_{\text{rms}} = \frac{f_m}{B} \sqrt{\frac{2P_s}{P_c}} \text{ Hz/Hz}$$

NOTE 1— The sideband response peaks do not usually present a steady oscilloscope image and some judgement must be used to estimate the time average of the maximum value.

NOTE 2 -The peak-to-peak deviation at modulation frequencies lower than the intermediate-frequency bandwidth can be estimated by observing the amount of jitter or thickening of the oscilloscope trace as a result of repeated carrier responses and relating it to the scale of the frequency sweep. The method is suitable only for comparatively high deviations and is usually limited by the frequency stability of the carrier.

5.2.6.3 Precautions — The following precautions shall be taken during measurement :

- a) Care should be taken to avoid overloading the mixer, particularly if several signals are present, since overloading may cause **inter-modulation**.
- b) The sweep repetition rate should be high enough to give continuous presentation if possible, but the frequency sweep rate should not exceed that for which the intermediate frequency and video circuits give their proper response.

APPENDIX A

(Clause 0.1)

TABLE OF MEASUREMENT METHOD CHARACTERISTICS

Clause Ref.	METHOD Description	USE	CARRIER	SIDE BAND FREQUENCY OF NOISE MEASUREMENT	NOISE WINDOW (WIDTH OF SIDE BAND MEASUREMENT)	TYPE OF NOISE: BACK-GROUND A.M. OR F.M.	NOISE LEVEL		A.M./ F.M. CORRELATION
							High	L o w	
5.1.1	Direct comparison with broadband noise source	General assessment of broadband amplifier	None		1 to 10 MHz (usually about 3 MHz)	Back-ground	Noise factor 30 dB	3 dB	No
5.1.2	Compares broad-band noise source by side-band suppression using tunable filter	Oscillator noise not immediately adjacent to carrier	1 to 5 mW	1 to 100 MHz (usually 10 to 70 MHz)	1 to 10 MHz (usually about 2 MHz)	Back-ground (and A.M.)	Carrier-to-noise ratio (referred to 1 Hz bandwidth) 90 dB	170 dB	No
5.1.3	Compares broad-band noise source by side-band suppression using two tunable filters	Oscillator noise not immediately adjacent to carrier	1 to 5 mW	1 to 100 MHz (usually 10 to 50 MHz)	1 to 10 MHz (usually 2 MHz)	A.M. and about F.M.	Carrier-to-noise ratio (referred to 1 Hz bandwidth) 90 dB	170 dB	Yes
5.1.4	Compares broad-band noise source by sideband suppression using balanced mixer	Oscillator noise not immediately adjacent to carrier	2 to 10 mW	1 to 100 MHz (usually 3 to 50 MHz)	1 kHz to 2 MHz dependent on sideband frequency	Back-ground A.M., F.M.	Carrier-to-noise ratio (1 Hz bandwidth) 90 dB	170 dB	Yes
5.2.1	Compares tube noise after detection with known signal at sideband frequency	Oscillator A.M. noise near carrier	1 to 5 mW	0 to 10 MHz (usually 10 Hz to 1 MHz)	10 Hz to 2 MHz dependent on sideband frequency	A.M. (and back-ground)	Carrier-to-noise ratio (1 Hz bandwidth) — dependent on sideband frequency near carrier	170 dB	No

5.2.2	Compares tube noise with rf signal having known modulation . F.M. noise obtained from "sideband slone " discriminator	Oscillator F.M. noise near carrier	2 to 10 mW	0 to 10 MHz (usually 10 Hz to 1 MHz)	10 Hz to 2 MHz (dependent on sideband frequency)	F.M. (and back-around)	Dependent on frequency near carrier	140 dB	No
5.2.3	Compares tube noise with broadband noise source of rf signal having F.M. noise obtained from two-path discriminator	Oscillator F.M. noise	10 to 20 mW	10 kHz to 10 MHz	1 kHz to 2 MHz	F.M.	Carrier-to-noise ratio (3 kHz band) 14 dB (deviation 3 kHz)	120 dB (deviation 10 kHz)	NO
5.2.4	Compares tube noise with rf signal having known modulation . F.M. noise obtained from frequency to Phase converter and phase detector	Oscillators A.M. and F.M. very near carrier	100 mW	50 Hz to 130 kHz	70 Hz to 1 kHz	A.M. (and back-ground) F.M. (and back-ground)	—	136 dB 107 dB (1 kHz from carrier)	Possible
5.2.5	Compares tube noise with rf signal having known A.M. or F.M. modulation by the use of a balanced phase discriminator	Amplifier modulation noise A.M. and F.M. c.w or pulsed	100 mW to 1000 mW c.w 1000 mW pulsed	0 Hz to 10 MHz usually 10 Hz to 1 MHz	10 Hz to 2 MHz	A.M. and F.M.	c.w signal-to-noise ratio 160 dB in 1 Hz bandwidth pulsed: 100 dB below main spectral lines in 1 Hz bandwidth		No
5.2.6	Measures modulation sideband power or undesired oscillation	Oscillators A.M. and F.M.	0.1 mW	-2 GHz	1 kHz to 2 MHz	A.M. and F.M. (and back-ground)	20 dB	90 dB	No

A P P E N D I X B

(Clause 3.2.2)

RELATIONSHIPS BETWEEN NOISE QUANTITIES

B-1. MODULATION NOISE

B-1.1 Modulation noise is generally related to the carrier level. Thus,

$$\left(\frac{C}{N}\right)_{\text{dB}} = 10 \log_{10} \frac{P_c}{P_n}$$

where

P_c = carrier power, and

P_n = noise power in a bandwidth B Hz.

If E is the amplitude of the unmodulated rf wave at carrier frequency F_c , ΔE is the amplitude of the modulating voltage at frequency F_m ,

$$M = \frac{\Delta E}{E} \text{ is the modulation index}$$

The ratio of the power in the carrier to the power in one sideband is

$$\left(\frac{C}{N}\right)_{\text{dB}} = 20 \log_{10} \frac{E}{\frac{\Delta E}{2}} \quad (1)$$

If modulation is noise rather than sinusoidal, the carrier-to-noise ratio is sometimes expressed as :

$$\left(\frac{C}{N}\right)_{\text{dB}} = 20 \log \frac{E}{\Delta E} \quad (2)$$

A common mistake arises from improper use of peak and rms values in the above formulae, and in not specifying whether one or both sidebands are being considered.

Formulae (1) and (2) give results differing by 3 dB in the case of sinusoidal modulation, but the results agree if both sidebands are considered (as they should be).

In the case of F.M. noise, it is common to refer to an effective (or rms) frequency deviation.

B-1.2 The modulation index of a frequency modulated carrier which is modulated by a discrete frequency to a certain frequency deviation is :

$$M = \frac{\Delta F}{f}$$

where

M = modulation index

ΔF = peak deviation in hertz of the carrier frequency F

f = modulating frequency in hertz.

The resulting first sidebands are located from the carrier by a frequency of $\pm f$, the second sidebands by $\pm 2f$, etc.

B-1.3 The amplitude of rf voltages in the carrier and sidebands vary as Bessel functions of the modulations index. The voltage amplitude of the carrier will vary as $J_0(M)$; the amplitude of the first sidebands as $J_1(M)$, etc. Therefore, the relative amplitudes of the power in the carrier and the power in one of the first order sidebands will be:

$$\left(\frac{C}{N}\right)_{\text{dB}} = 20 \log_{10} \frac{J_0(M)}{J_1(M)} \quad (3)$$

Where the modulation index is small:

$$J_0(M) \cong 1.0$$

$$J_1(M) \cong \frac{M}{2} \cong \frac{\Delta F}{2f}$$

Then (3) becomes:

$$\left(\frac{C}{N}\right)_{\text{dB}} = 20 \log_{10} \frac{2f}{\Delta F} \quad (4)$$

Equation (4) gives the relationship between the carrier and the power in one of the first order sidebands for small modulation indices.

It should be noted that the carrier-to-sideband power ratio varies both with peak deviation ΔF and the distance away from the carrier f .

With this relationship, it is easy to correlate a noise power measured as so many decibels below the carrier in stated bandwidth, at any place in the spectrum with an equivalent power which would be generated by a discrete modulation at the same place in the spectrum.

B-1.4 It is usual to define F.M. noise in terms of rms frequency deviation in both sidebands. In this case, we have:

$$\left(\frac{C}{N}\right)_{\text{dB}} = 20 \log_{10} \frac{f}{\Delta F_{\text{rms}}} \quad (\text{two sidebands}) \quad (5)$$

If A_1 is the decibel carrier-to-noise power ratio in a bandwidth of measurement B_1 Hz, the A_2 decibel carrier-to-noise power ratio in a B_2 Hz bandwidth is given by:

$$A_2 = A_1 + 10 \log \frac{B_1}{B_2}$$

IS : 6134 (Part V) - 1980

If ΔF_1 is the deviation of the carrier frequency in a bandwidth of measurement B_1 Hz, the corresponding ΔF_2 deviation of the carrier frequency in a B_2 Hz bandwidth is:

$$\Delta F_2 = \Delta F_1 \sqrt{\frac{B_2}{B_1}}$$

NOTE — These relations apply only when measured noise appears as a continuous spectrum. They do not apply when coherent spectral lines are present due for instance to power supply ripple. Considering noise measurements under pulsed conditions the same relationships apply, but it is possible to relate the noise power to three different values of the carrier power: power of the unmodulated carrier P_o (peak power), mean (or average) transmitted power P_m , or power in one of the main spectral lines P_1 . The following relationship applies:

$$P_1 = P_m \left(\frac{t}{T} \right) = P_o \left(\frac{t}{T} \right)^2$$

where

t = pulse duration, and

T = pulse repetition period.

INDIAN STANDARDS

ON

ELECTRON TUBES

IS:

- 1885 (Part **IV/Sec 1**)-1973 Electrotechnical vocabulary: Part IV Electron tubes, Section 1
Common terms (*first revision*)
- 1885 (Part **IV/Sec 2**)-1973 Electrotechnical vocabulary: Part IV Electron tubes, Section 2
X-ray tubes (*first revision*)
- 1885 (Part **IV/Sec 3**)-1970 Electrotechnical vocabulary: Part IV Electron tubes, Section 3
Microwave tubes
- 1885 (Part **IV/Sec 4**)-1970 Electrotechnical vocabulary: Part IV Electron tubes, Section 4
Cathode-ray tubes
- 1885 (Part **IV/Sec 5**)-1972 Electrotechnical vocabulary: Part IV Electron tubes, Section 5
Pulse terms
- 1885 (Part **IV/Sec 6**)-1972 Electrotechnical vocabulary: Part IV Electron tubes, Section 6
Noise in microwave tubes
- 1883 (Part **IV/Sec 7**)-1973 Electrotechnical vocabulary: Part IV Electron tubes, Section 7
Camera tubes
- 1885 (Part **IV/Sec 8**)-1973 Electrotechnical vocabulary: Part IV Electron tubes, Section 8
Photosensitive devices
- 2032 (Part IX)-1969 Graphical symbols used in electrotechnology: Part IX Electron tubes (other than microwave tubes)
- 2032 (Part **XIII**)-1971 Graphical symbols used in electrotechnology: Part XIII Microwave tubes
- 2597 (Part **I**)-1964 Code of practice for the use of electron tubes: Part I Commercial receiving tubes
- 2597 (Part **II**)-1967 Code of practice for use of electron tubes: Part II Special quality receiving tubes
- 2597 (Part **III**)-1969 Code of practice for the use of electron tubes: Part III Transmitting and industrial tubes
- 2597 (Part **IV**)-1970 Code of practice for the use of electron tubes: Part IV Cathode-ray tubes
- 2597 (Part **V**)-1971 Code of practice for the use of electron tubes: Part V Rectifiers and thyristors
- 2612-1965 Recommendation for type approval and sampling procedures for electronic components
- 2684 (Part **I**)-1972 Dimensions of electron tubes: Part I Miniature '9-pin noval' type (*first revision*)
- 2684 (Part **II**)-1972 Dimensions of electron tubes: Part II Miniature '7-pin type' (*first revision*)
- 2684 (Part **III**)-1971 Dimensions of electron tubes: Part III Octal base type
- 2684 (Part **IV**)-1971 Dimensions of electron tubes: Part IV Magnoval base type
- 2684 (Part **V**)-1972 Dimensions of electron tubes: Part V Localt base type
- 3154-1965 Specification for X-ray tubes, diagnostic type
- 4096-1973 Method of measurement of optical focal spot size of X-ray tubes (*first revision*)
- 4147-1967 Method of measurements on conventional receiving electron tubes
- 4579-1968 Method of measurements on television picture tubes
- 4697-1968 Method of measurements on Geiger Muller counter tubes
- 5323-1969 Letter symbols and abbreviations for electron tubes
- 5627-1970 Methods of measurements on cathode-ray display tubes
- 5840 (Part **I**)-1970 Dimensions of cathode-ray tubes : Part I Tube outlines
- 5840 (Part **II**)-1970 Dimensions of cathode-ray tubes : Part II Bases
- 5840 (Part **III**)-1970 Dimensions of cathode-ray tubes : Part III EHT Terminals
- 6134 (Part **I**)-1978 Methods of measurements on microwave tubes: Part I Common to all microwave tubes [Superseding IS: 6134 (Part **I/Sec 1**) and IS : 6134 (Part **I/Sec 2**)]
- 6134 (Part **II**)-1973 Methods of measurement on microwave tubes : Part II Oscillator tubes

IS:

- 6134 (Part **III**)-1973 Methods of measurement on microwave tubes : Part III **Amplifier** tubes
- 6134 (Part **IV**)-1977 Methods of measurement on microwave tubes : Part IV Magnets
- 6134 (Part **W**)-1980 Methods of measurement on microwave tubes : Part V Parasitic noise
- 6136-1971 Basic requirements for cathode-ray tubes
- 6214-1971 Phosphors for cathode-ray tube
- 6567-1972 Radiation protection for an X-ray tube in a protective tube-housing operating between **10 kV** and **400 kV**
- 6568-1972 Implosion protection for TV picture tubes
- 6576-1972 Methods of measurements on gas filled cold cathode indicator tubes
- 6577-1972 Methods of measurements on gas filled cold cathode voltage stabilizing and voltage reference tubes
- 6757-1972 Dimensions for high tension cable terminations for X-ray tubes
- 6758-1972 Dimensions for high tension receptacles for X-ray tubes
- 7012-1973 Specification for X-ray tube shield
- 7144-1973 Methods of measurements on camera tubes
- 7146 (Part **Q**)-1973 Methods of measurements on **photosensitive** devices: Part I Basic considerations
- 7146 (Part **II**)-1974 Methods of measurements on photosensitive devices: Part II **Photo-tubes**
- 7146 (Part **III**)-1974 Methods of measurements of photosensitive devices: Part III Photo-conductive cells for use in the visible spectrum
- 7146 (**Part IV**)-1974 Methods of measurements on photosensitive devices: Part IV Photo-multipliers
- 8319 (Part **I**)-1977 Dimensions of indicator tubes : Part **I** Tube type 1
- 8319 (Part **II**)-1977 Dimensions of indicator tubes : Part II Tube type 2
- 8319 (Part **III**)-1977 Dimensions of indicator tubes : Part III Tube type 3
- 8319 (Part **IV**)-1979 Dimensions of indicator tubes : Part IV Tube **type 4**
- 8441**-1977 Methods of measurements on incidental X-radiation from-electron tubes
- 9383 (Part **I**)-1979 Dimensions of sign indicator tubes : Part I **Tube type 1**
- 9383 (Part **II**)-1979 Dimensions of sign indicator tubes : Part II Tube type 2

INTERNATIONAL SYSTEM OF UNITS (SI UNITS)

Base Units

<i>Quantity</i>	<i>Unit</i>	<i>Symbol</i>
Length	metre	m
Mass	kilogram	kg
Time	second	s
Electric current	ampere	A
Thermodynamic temperature	kelvin	K
Luminous intensity	candela	cd
Amount of substance	mole	mol

Supplementary Units

<i>Quantity</i>	<i>Unit</i>	<i>Symbol</i>
Plane angle	radian	rad
Solid angle	steradian	sr

Derived Units

<i>Quantity</i>	<i>Unit</i>	<i>Symbol</i>	<i>Definition</i>
Force	newton	N	1 N = 1 kg.m/s ²
Energy	joule	J	1 J = 1 N.m
Power	watt	W	1 W = 1 J/s
Flux	weber	Wb	1 Wb = 1 V.s
Flux density	tesla	T	1 T = 1 Wb/m ²
Frequency	hertz	Hz	1 Hz = 1 c/s (s ⁻¹)
Electric conductance	siemens	S	1 S = 1 A/V
Electromotive force	volt	V	1 V = 1 W/A
Pressure, stress	pascal	Pa	1 Pa = 1 N/m ²

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